

Islamic University-Gaza

Deanery of Graduate Science

Faculty of Science

Department of Physics

# **Transmission Lines**

## **and**

# **Schottky Diode**

by

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”A thesis submitted in partial fulfillment of the requirements for the degree  
of master of science in Physics faculty of science Islamic University of  
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يَا أَيُّهَا الَّذِينَ آمَنُوا اذْكُرُوا اللَّهَ ذِكْرًا كَثِيرًا (41)  
وَسَبِّحُوا بِحَمْدِهِ وَاصْبِرُوا (42) هُوَ الَّذِي يُصَلِّيْ عَلَىكُمْ  
وَمَلَكَكُمْ لِنُجْحِكُمْ مِنَ الظُّلُمَاتِ إِلَى النُّورِ وَكَانَ بِالْمُؤْمِنِينَ حَمِيمًا (43)  
تَحِيَّةٌ مِّمَّنْ يَلْقَوْنَ سُبْحًا وَعَدْلٌ مِّمَّنْ جَاءَكَ يَأْتِيهِمُ النَّبِيُّ  
إِنَّا أَرْسَلْنَاكَ شَهِيدًا وَنَذِيرًا (45) وَرَاعِيْنَا إِلَى اللَّهِ بِإِذْنِهِ  
وَسِرَاجًا مُنِيرًا (46) وَبَشِّرِ الْمُؤْمِنِينَ بِالْجَنَّةِ فَضْلًا كَثِيرًا (47)  
وَلَا تُطِيعُوا الْكَافِرِينَ وَالْمُنَافِقِينَ دَعَاؤُهُمْ  
وَتَوَكَّلْ عَلَى اللَّهِ وَكَفَى بِاللَّهِ وَكِيلًا (48)

الأحزاب الآية 41 - 48

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## ملخص

### خطوط النقل وثنائي شوتكى

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ديسمبر ٢٠١٠ م - محرم ١٤٣٢ هـ

إن هذه الأطروحة تعالج خطوط النقل والخطية وثنائي شوتكى الغير خطى حيث حدث مؤخرا تقدم كبير في صناعة أشباه الموصلات على مستوى العالم ونظرا لأن ثنائي شوتكى يعد نموذجا مميزا للثنائيات

فقد قمت بدراسة لجوانب مختلفة من هذا الثنائي من حيث تركيبه واليه عمله. قمت باستخدام علاقة غير خطية في هذا الثنائي وكذلك قمت باشتقاق معادلة رياضية في صيغتها العامة مشابهة لمعادلة فان دير بول المعروفة في عالم المتذبذبات .

وكذلك قمت باستخدام برنامج محاكاة لرسم علاقة بين فرق الجهد والزمن التي أظهرت كون الثنائي النفقي الرنينى متذبذب إشارة متغيرة فى القيمة مما يساعد فى تقوية الأمواج المنقادة التي تنقل الإشارات وقد اعتمدت على خاصية تسمى المقاومة السالبة التي تظهر عندما يزيد فرق الجهد يقل التيار فى حدود معينة وهذه الخاصية تؤدي الى التوفير فى الطاقة .

وختاما ارجوا أن يتاح المجال لإنتاج أشياء مفيدة من هذا الثنائي

# Abstract

In this thesis we used a nonlinear relation between current and voltage for schottky resonant tunneling diode and obtained a general form of equation similar to Van Der Pol equation for an oscillator.

A Resonant tunneling diode(RTD) has a negative differential resistance that means when the voltage increases the current decreases. This property is very useful for oscillators manufacture.

Also I studied some kinds of nonlinear transmission lines to show that it can be used in oscillators and show that it can reshape the sinusoidal signals to other shapes by using OrCad and mathematica programming .



# Introduction

Information transmission rates and operating frequencies of electronic systems have increased dramatically since the birth of electrical telegraphy in the year 1836 and communication links have progressed from Morse code transmitted at a few symbols per second on a pair of wires, through telephone with ( 4 kHz) for voice transmission in the year(1876), culminating in fibre optic transmission system carrying 150000 conversation with(10 GB/s) data rates .

Broadcast frequencies have progressed from the 800 kHz for Marconi's in the year 1901 through 500 kHz amplitude radio to 12 GHz direct broadcast satellite television, demand for communications capacity is driven by growing use of computers, cable television, portable telephones and facsimile machines, a change in high speed electronics will permit 100-GB/s data transmission on optical fibre, wave propagation in unbounded media is said to be unguided and exist through all space at EM (electromagnetic) energy associated with waves such as radio and TV (television) waves . Another means of transmitting power and information are guided structures, typical examples of such structures are transmission lines and waveguides. Transmission lines are commonly used in power distribution at low frequencies for the propagation of energy from source to the load and in communication at high frequency .

Diodes are fundamental electronic building blocks. Their ability to restrict current flow to one direction is a critical property, diodes are used in every electronic circuit manufactured, from the smallest power supply to the largest industrial process controller.

In recent years, mobile communication equipment such as digital cellular phones and high-speed data communication equipment have been required to be small, thin

and lightweight, low power consumption, a high frequency, and multi bands and diodes did this.

A Schottky diode is a special type of diode with a very low forward-voltage drop.

When current flows through a diode, it has some internal resistance to that current flow, which causes a small voltage drop across the diode terminals, a normal diode has between 0.7-1.7 volt drops, while a Schottky diode voltage drop is between approximately 0.15-0.45 . This lower voltage drop translates into higher system efficiency . Schottky diodes are semiconductor devices which have a metal-semiconductor transition as their basic structure and whose basic electronic properties are defined by this transition. Schottky diode are used in the high-frequency module are required to be small. Schottky diodes have been used for many years in the semiconductor industry, for example in electronic systems such as amplifiers, receivers, control and guidance systems, power and signal monitors, and as rectifiers, also in commercial applications include radiation detectors, imaging devices, and wired and wireless communications products. Typically, a Schottky diode consists of a metal layer connected to a doped semiconductor layer, the Schottky barrier is formed at the juncture of the metal and the semiconductor, this leads to a low breakdown voltage .

In this thesis, I studied the behavior of nonlinear transmission line and showed that it change the shape of the signals and it can be used as an oscillators by using OrCad and mathematica programs. I deduced transmission line equations and their parameters in the first chapter.

In the second chapter, I studied the difference between liner and nonlinear transmission lines .In the third chapter, I presented a study about Smith Chart and how to used it in finding transmission line parameter. In the fourth chapter, I considered resonant tunneling diodes and used nonlinear relations in oscillators. In the last chapter, I presented a study about Schottky diode and my work in Schottky diode to show that it can be used as an oscillator .

# Chapter 1

## Transmission Lines

### 1.1 Transmission Lines

Transmission line basically[1] consists of two or more parallel conductors used to connect a source to a load. The source may be hydroelectric generator, transmitter or oscillator, the load may be a factory, an antenna or an oscilloscope, respectively. Transmission line includes coaxial cable, a two-wire line a parallel and microstrip line as shown in Fig.(1.1)

These lines are consist of two conductors in parallel. Coaxial cable are used in electrical laboratories and in connecting TV sets to TV antennas's while microstrip lines are used in integrated circuit.

Transmission line problem are solved by using electrical circuit theory.

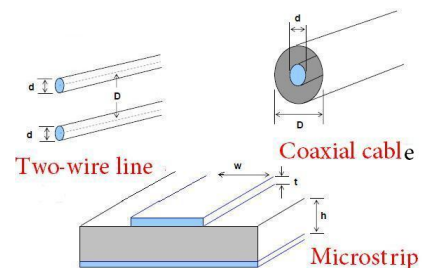


Figure 1.1: Schematic structure of there types of transmission lines

### 1.2 Transmission Line Parameters

Its convenient to describe transmission line in terms of its line parameters which are resistance per unit length  $R$ , inductance per unit length  $L$ , conductance per unit length  $G$  and capacitance per unit length  $C$ . The parameters  $R$ ,  $L$ ,  $G$  and  $C$  are distributed uniformly along the length of the line.  $G \neq 1/R$ ,  $R$  is AC resistance

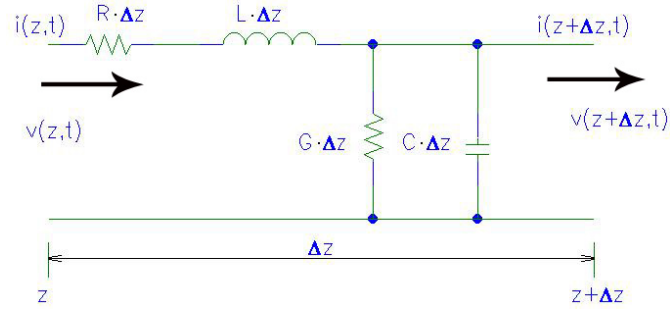


Figure 1.2: Equivalent circuit of transmission line

per unit length of the conductors and  $G$  is the conductance per unit length due to dielectric medium separating the conductors.

### 1.3 Transmission Line Equations

Two conductor transmission line supports a TEM (transverse electromagnetic) wave that is the electric and magnetic fields on the line are transverse to the direction of wave propagation. An important property of TEM wave is that the fields  $E$  and  $H$  are uniquely related to voltage  $V$  and current  $I$  respectively.  $V = -\int \underline{E} \cdot d\underline{\ell}$

$$I = \oint \underline{H} \cdot d\underline{\ell}$$

In view of this, we will use circuit quantities  $V$  and  $I$  in solving the transmission line problem instead of solving field quantities  $E$  and  $H$ . Let us examine an incremental portion of length  $\Delta z$  of a two conductor transmission from Fig.(1.2) using kirchhoff's voltage law.[1,2]

$$V(z, t) = R\Delta z I(z, t) + L\Delta z \frac{\partial I(z, t)}{\partial t} + V(z + \Delta z, t) \quad (1.1)$$

Divided by  $\Delta z$

$$-\frac{V(z + \Delta z, t) - V(z, t)}{\Delta z} = RI(z, t) + L\frac{\partial I(z, t)}{\partial t} \quad (1.2)$$

$\Delta z$  is small so that Eq.(1.2) can be written as

$$-\frac{\partial V(z, t)}{\partial z} = RI(z, t) + L\frac{\partial I(z, t)}{\partial t} \quad (1.3)$$

Similarly applying kirchhoff's current law in Fig.(1.2) gives

$$I(z, t) = I(z + \Delta z) + \Delta I = I(z + \Delta z) + G\Delta z V(z + \Delta z, t) + C\Delta z \frac{\partial V}{\partial t}(z + \Delta z, t) \quad (1.4)$$

Dividing by  $\Delta z$ , as  $\Delta z$  is small so Eq.(1.4) can be written as

$$-\frac{\partial I(z, t)}{\partial z} = GV(z, t) + C\frac{\partial V(z, t)}{\partial t} \quad (1.5)$$

Assume harmonic time dependence so that the solution can be written as

$$V(z, t) = R_e[V_s e^{j\omega t}] \quad (1.6)$$

$$I(z, t) = R_e[I_s e^{j\omega t}] \quad (1.7)$$

where  $V_s$  and  $I_s$  are phasors form of  $V(z, t)$  and  $I(z, t)$ , respectively.

Eqs. (1.3) and (1.5) can be written as [see appendix A]

$$-\frac{\partial V_s}{\partial z} = (R + j\omega L)I_s \quad (1.8)$$

$$-\frac{\partial I_s}{\partial z} = (G + j\omega C)V_s \quad (1.9)$$

Takeing the second derivatives to Eqs.(1.8) and (1.9) and substitute from one to anther we get

$$\frac{\partial^2 V_s}{\partial z^2} = (R + j\omega L)(G + j\omega C)V_s \quad (1.10)$$

$$\frac{\partial^2 V_s}{\partial z^2} - \gamma^2 V_s = 0 \quad (1.11)$$

Similarly for current we get

$$\frac{\partial^2 I_s}{\partial z^2} - \gamma^2 I_s = 0 \quad (1.12)$$

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \quad (1.13)$$

where  $\gamma$  is the propagation constant,  $\alpha$  is the attenuation constant, the phase constant  $\beta$ , is given by  $\beta = \frac{2\pi}{\lambda} = \frac{w}{u}$ , wave length  $(\lambda) = \frac{2\pi}{\beta}$ , wave velocity  $(u) = \frac{w}{\beta}$ , and angular velocity  $(\omega) = 2\pi f$

The solution for Eqs .(1.11) and (1.12)

are

$$V_s(z) = V_o^+ e^{-\gamma z} + V_o^- e^{\gamma z} \quad (1.14)$$

$$I_s(z) = I_o^+ e^{-\gamma z} + I_o^- e^{\gamma z} \quad (1.15)$$

where the first term for  $+z$  direction and the second term for  $-z$  direction

where

$V_o^+, I_o^+, V_o^-$  and  $I_o^-$ , are wave amplitude along  $+z$  and  $-z$  directions.

## 1.4 Characteristic Impedance

Characteristic impedance  $Z_o$  of the transmission line is the ratio of positively voltage wave to current wave at any point of the line.[2,3]

$$Z_o = \frac{V_o^+}{I_o^+} = \frac{V_o^-}{I_o^-} \quad (1.16)$$

Differentiate Eqs. (1.14 and 1.15) with respect to  $z$ , and equating them with Eqs.(1.8,1.9) and equating coefficient of  $e^{\gamma z}, e^{-\gamma z}$

We get [See Appendix B ]

$$Z_o = \frac{R + jwL}{\gamma} = \frac{\gamma}{G + jwC} \quad (1.17)$$

$$Z_o = \sqrt{\frac{R + jwL}{G + jwC}} = R_o + jX_o \quad (1.18)$$

$R_o$  and  $X_o$  are real and imaginary parts of  $Z_o$

The propagation constant  $\gamma$  and characteristic impedance  $Z_o$  are important properties of the transmission line because they both depend on the line parameters  $R$ ,  $L$ ,  $G$  and  $C$  and the frequency of operation, the reciprocal of  $Z_o$  is the characteristic admittance  $Y_o$ , that is

$$Y_o = \frac{1}{Z_o} \quad (1.19)$$

## 1.5 Lossless Line

The line is said to be lossless if conductors of the line are perfect, then Its conductance is

$$\sigma_c = \infty$$

and the dielectric medium separating them is lossless ( $\sigma = 0$ ) and the condition to be lossless that [1]

$$R = 0 = G$$

$$\alpha = 0$$

from Eqs.(1.13) and (1.18) we get

$$\gamma = j\beta = j\omega\sqrt{LC}$$

$$\beta = \omega\sqrt{LC}$$

$$X_o = 0$$

$$\mathbf{Z}_o = \mathbf{R}_o = \sqrt{\frac{\mathbf{L}}{\mathbf{C}}} \quad (1.20)$$

$$\mathbf{u} = \frac{\omega}{\beta} = \frac{1}{\sqrt{\mathbf{LC}}} = \mathbf{f}\lambda \quad (1.21)$$

## 1.6 Distortionless Line

Distortionless line; this line does not affect on the amplitude of the wave such that  $\alpha$  (attenuation constant), is independent of frequency and  $\beta$  linearly dependent on frequency and the condition for this line is that[1,2]

$$\frac{R}{L} = \frac{G}{C} \quad (1.22)$$

$$\gamma = \sqrt{RG(1 + \frac{j\omega L}{R})(1 + \frac{j\omega C}{G})} \quad (1.23)$$

$$= \sqrt{RG}(1 + \frac{j\omega C}{G}) = \alpha + j\beta \quad (1.24)$$

or

$$\alpha = \sqrt{RG}, \beta = \omega\sqrt{LC} \quad (1.25)$$

from Eq. (1.18) we get

$$Z_o = \sqrt{\frac{R(1 + j\omega\frac{L}{R})}{G(1 + j\omega\frac{C}{G})}} \quad (1.26)$$

$$Z_o = \sqrt{\frac{R}{G}} = \sqrt{\frac{L}{C}}, X_o = 0 \quad (1.27)$$



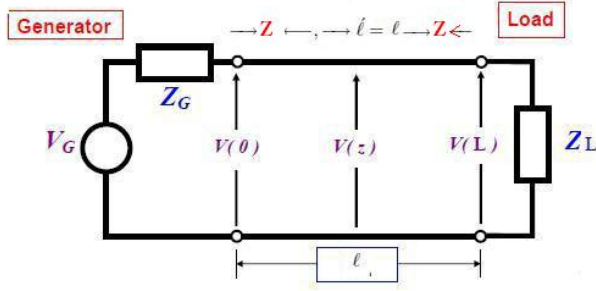


Figure 1.3: Input impedance at the load

## 1.7 Input Impedance:

Consider a transmission line of length  $\ell$ , characterized by  $\gamma$  and  $Z_o$  connected to the load  $Z_L$  as shown in Fig.(1.3), the generator sees the line with the load as an input impedance  $Z_{in}[1,2]$

$\ell$  and  $z$  are used for length,  $L$  for load, and  $Z$  for impedance, let the transmission line extend from  $z = 0$  at the generator to  $z = \ell$  at the load, we need current and voltage equations

$$V_s(z) = V_o^+ e^{-\gamma z} + V_o^- e^{\gamma z} \quad (1.28)$$

$$I_s(z) = \frac{V_o^+}{Z_o} e^{-\gamma z} - \frac{V_o^-}{Z_o} e^{\gamma z} \quad (1.29)$$

To find  $V_o^-$ ,  $V_o^+$  terminal condition must be given.

Assume the condition

$$V_o = V(z = 0)$$

$$I_o = I(z = 0)$$

Substituting these into Eqs. (1.28) and (1.29) we get

$$V_o = V_o^+ + V_o^- \quad (1.30)$$

$$Z_o I_o = V_o^+ - V_o^- \quad (1.31)$$

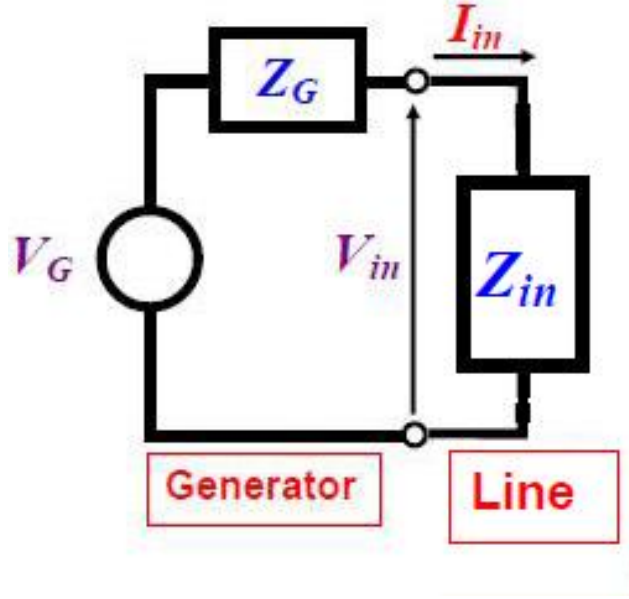


Figure 1.4: The voltage at the input

Adding and subtracting Eqs.(1.30) and (1.31) we get

$$V_o^+ = \frac{1}{2}(V_o + Z_o I_o) \quad (1.32)$$

$$V_o^- = \frac{1}{2}(V_o - Z_o I_o) \quad (1.33)$$

If the input impedance at the input terminal is  $Z_{in}$ , then the input voltage  $V_{in}$  and the input current  $I_{in}$  is given by

$$V_{in} = I_{in} Z_{in} \quad (1.34)$$

$$I_{in} = \frac{V_G}{Z_{in} + Z_G} \quad (1.35)$$

$$V_{in} = \frac{Z_{in}}{Z_{in} + z_G} V_G \quad (1.36)$$

If we take the condition at the load (L)

$$V_L = V(z = \ell)$$

$$I_L = I(z = \ell)$$

and substituting into Eqs. (1.28 and 1.29) we get

$$V_L = V_o^+ e^{-\gamma\ell} + V_o^- e^{\gamma\ell} \quad (1.37)$$

$$Z_o I_L = V_o^+ e^{-\gamma\ell} - V_o^- e^{\gamma\ell} \quad (1.38)$$

From this ,we get

$$V_o^+ = \frac{1}{2}(V_L + Z_o I_L) e^{\gamma\ell} \quad (1.39)$$

$$V_o^- = \frac{1}{2}(V_L - Z_o I_L) e^{-\gamma\ell} \quad (1.40)$$

The input impedance

$$Z_{in} = \frac{V_s(z)}{I_s(z)} \quad (1.41)$$

From Eqs.(1.28 and 1.29) and at  $z = 0$  we get

$$Z_{in} = \frac{V_s(z)}{I_s(z)} = \frac{Z_o(V_o^+ + V_o^-)}{(V_o^+ - V_o^-)} \quad (1.42)$$

Substituting Eqs.(1.39 and 1.40) into Eq.(1.42) and utilizing the fact that

$$\frac{e^{\gamma\ell} + e^{-\gamma\ell}}{2} = \cosh \gamma\ell \quad (1.43)$$

$$\frac{e^{\gamma\ell} - e^{-\gamma\ell}}{2} = \sinh \gamma\ell \quad (1.44)$$

or

$$\tanh \gamma\ell = \frac{\sinh \gamma\ell}{\cosh \gamma\ell} = \frac{e^{\gamma\ell} - e^{-\gamma\ell}}{e^{\gamma\ell} + e^{-\gamma\ell}} \quad (1.45)$$

$$Z_{in} = Z_o \left[ \frac{Z_L + Z_o \tanh \gamma \ell}{Z_o + Z_L \tanh \gamma \ell} \right] \quad (1.46)$$

Eq.(1.46) is for Lossy Line [1]

$Z_{in}$  is input impedance at generation end, to find  $Z_{in}$  at distance  $\ell$  from the load we replace  $\ell$  by  $\ell$

For lossless line, the conductance of the line are perfect  $\sigma_c \approx \infty$

$\alpha = zero, \gamma = j\beta, \tanh j\beta\ell = j \tan \beta\ell, Z_o = R_o$

$$Z_{in} = Z_o \left[ \frac{Z_L + jZ_o \tan \beta\ell}{Z_o + jZ_L \tan \beta\ell} \right] \quad (1.47)$$

Eq.(1.47) Shows that the input impedance varies periodically with distance  $\ell$  from the load,  $\beta\ell$  is usually referred to the electrical length of the line and can be expressed in degree or radians.

Voltage reflection coefficient  $\Gamma_L$  (at the load) is the ratio of the voltage reflected wave to the incident wave that is

$$\Gamma_L = \frac{V_o^- e^{\gamma\ell}}{V_o^+ e^{-\gamma\ell}} \quad (1.48)$$

Substituting Eqs.(1.39 and 1.40) into Eq.(1.42)

and we have :

$$V_L = I_L Z_L \quad (1.49)$$

we get

$$\Gamma_L = \frac{Z_L - Z_o}{Z_L + Z_o} \quad (1.50)$$

From

$$\Gamma_L = \frac{V_o^-}{V_o^+} e^{2\gamma\ell} \quad (1.51)$$

At any point z

$$z = \ell - \ell' \quad (1.52)$$

$$\Gamma(z) = \frac{V_o^-}{V_o^+} e^{2\gamma z} \quad (1.53)$$

$$= \frac{V_o^-}{V_o^+} e^{2\gamma(\ell - \ell')} \quad (1.54)$$

$$= \Gamma_L e^{-2\gamma \ell'} \quad (1.55)$$

Current reflection coefficient at any point is negative of the voltage reflection coefficient at this point.

$$\frac{I_o^- e^{\gamma \ell}}{I_o^+ e^{-\gamma \ell}} = -\Gamma_L \quad (1.56)$$

Standing wave ratio(S)

$$S = \frac{V_{max}}{V_{min}} = \frac{I_{max}}{I_{min}} \quad (1.57)$$

$$= \frac{1 + |\Gamma_L|}{1 - |\Gamma_L|} \quad (1.58)$$

## Chapter 2

# Non-Linear Transmission Lines(NLTL)

### linear system

In this system the output is related to the input, when the input is a wave of certain frequency then the output is of the same frequency, the linear system is used in wire and wireless communication.

### 2.1 Nonlinear Transmission Line Theory

Definition in mathematics, a nonlinear system is a system which is not linear, that is, a system which does not satisfy the superposition principle, or whose output is not directly proportional to its input. Less technically, a nonlinear system is any problem where the variable(s) to be solved for cannot be written as a linear combination of independent components.[4]

Most of real life problems involve nonlinear systems, the nonlinear system are very hard to solve explicitly but qualitative and numerical techniques may help to give some information on the behavior of the solutions . In the nonlinear system when the input is a wave of certain frequency then the output is a group of other frequencies . The NLTL has three fundamental and quantifiable characteristics just as any non ideal transmission line. These are nonlinearity, dispersion, and dissipation.

## Dispersion

Dispersion is a variation in phase velocity with frequency. At low frequencies, loss is dominated by coplanar waveguide (CPW) resistivity. At high frequencies, loss is dominated by diode series resistance.

## Nonlinearity

Diodes present two sources of nonlinearity: conductive and reactive. The conductive nonlinearity is evident in the  $I(V)$  (current and voltage) curves and the reactive nonlinearity is evident in the  $C(V)$  (capacitance and voltage) curves .

## Dissipation

There are two main sources of dissipation in an NLTL. These are diode series resistance and metallic losses. Diode losses arise from the nonzero contact and bulk resistances of the structure while metallic losses arise from the geometry and finite conductivity of the CPW. Another source of loss is radiation, where some portion of the propagating energy is coupled into the substrate; but this loss mechanism is much less significant in an NLTL than the other two.

## 2.2 Linearization Technique

Recall that only the solutions of linear systems may be found explicitly. The problem is that in general real life problems may only be modeled by nonlinear systems. In this case, we only know how to describe the solutions globally (via nullclines). What happens around an equilibrium point remains a mystery so far. The main idea is to approximate a nonlinear system by a linear one (around the equilibrium point = 0). Of course, we do hope that the behavior of the solutions of the linear system will be the same as the nonlinear one. This is the case most of the time .

Example. Consider the Van der Pol equation

$$\ddot{x} - \mu(1 - x^2)\dot{x} + x = 0. \quad (2.1)$$

where  $x$  is the voltage and  $\mu$  is a factor (see chapter 4). This is a nonlinear equation, to translate this equation into a linear system.

we set

$$\dot{x} = y, \dot{y} = -x + \mu(1 - x^2)y \quad (2.2)$$

## 2.3 Electronic Devices

- An Electronic Mixer

An electronic mixer is a device that combines two or more electronic signals into one composite output signal. There are two basic types of mixers. Additive mixers add two signals together, and are used for such applications as audio mixing. Multiplying mixers multiply the signals together, and produce an output containing both original signals, and new signals that have the sum and difference of the frequency of the original signals.

- Electronic Oscillator

An electronic oscillator is an electronic circuit that produces a repetitive electronic signal, often a sine wave or a square wave, a low-frequency oscillator (LFO) is an electronic oscillator that generates an AC waveform at a frequency below  $\approx 20Hz$ .

Oscillators designed to produce a high-power AC output from a DC supply are usually called inverters.

The waveform generators which are used to generate pure sinusoidal waveforms of fixed amplitude and frequency are called oscillators, these oscillators are basic to modern communications because they provide the carriers for transmission of sound and picture

- Harmonic Oscillator

The harmonic, or linear, oscillator produces a sinusoidal output. The basic form of a harmonic oscillator is an electronic amplifier with the output attached to an electronic filter, and the output of the filter attached to the input of the amplifier, in a feedback loop. When the power supply to the amplifier is first



switched on, the amplifier's output consists only of noise. The noise travels around the loop, being filtered and re-amplified until it increasingly resembles the desired signal.

A piezoelectric crystal (commonly quartz) may take the place of the filter to stabilize the frequency of oscillation, resulting in a crystal oscillator.

There are many ways to implement harmonic oscillators, because there are different ways to amplify and filter.

- Detector

A detector is a device capable of registering a specific substance or physical phenomenon.

# Chapter 3

## Smith Chart

### 3.1 Graphical Methods

Transmission line problems often involve manipulations with complex numbers, making the time and effort required for a solution several times greater than that needed for a similar sequence of operations on real numbers. One means of reducing the effort without seriously affecting the accuracy is by using transmission-line charts. Perhaps the most widely used one is the Smith Chart.[1,2]

Basically, this diagram shows curves of constant resistance and constant reactance, these may represent either an input impedance or a load impedance. "An indication of location along the line is also provided, usually in terms of the fraction of a wavelength from a voltage maximum or minimum". Although they are not specifically shown on the chart, the standing-wave ratio and the magnitude and angle of the reflection coefficient are very quickly determined.

As a matter of fact, the diagram is constructed within a circle of unit radius, using polar coordinates, with radius variable  $|\Gamma|$  and counterclockwise angle variable  $\phi$ , where  $\Gamma = |\Gamma|e^{j\phi}$  Fig.(3.1) shows this circle. Since  $\Gamma \leq 1$ , all our information must lie on or within the unit circle.

The reflection coefficient itself will not be plotted on the final chart, for these additional contours would make the chart very difficult to read.

The basic relationship upon which the chart is constructed is

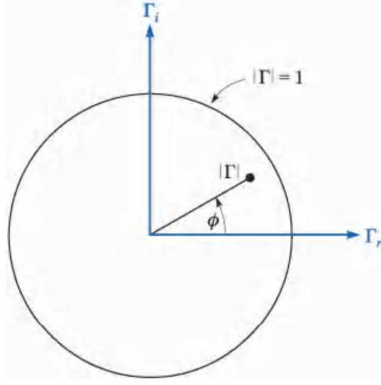


Figure 3.1: Complex  $\Gamma$  plain

$$\Gamma = \frac{Z_L - Z_o}{Z_L + Z_o} \quad (3.1)$$

The impedances which we plot on the chart will be normalized with respect to the characteristic impedance. Let us identify the normalized load impedance as  $z_L$ ,

$$z_L = r + jx = \frac{Z_L}{Z_o} = \frac{R_L + jX_L}{Z_o} \quad (3.2)$$

and thus

$$\Gamma = \frac{z_L - 1}{z_L + 1} \quad (3.3)$$

or

$$z_L = \frac{1 + \Gamma}{1 - \Gamma} \quad (3.4)$$

Let us now select  $\Gamma_r$  and  $\Gamma_i$ , as the real and imaginary parts of  $\Gamma$  in cartesian coordinates .

$$\Gamma = \Gamma_r + j\Gamma_i \quad (3.5)$$

thus

$$z_L = r + jx = \frac{1 + \Gamma_r + j\Gamma_i}{1 - \Gamma_r - j\Gamma_i} \quad (3.6)$$

We can write the real and imaginary parts of this equation as

$$r = \frac{1 - \Gamma_r^2 - \Gamma_i^2}{(1 - \Gamma_r)^2 + \Gamma_i^2} \quad (3.7)$$

$$x = \frac{2\Gamma_i}{(1 - \Gamma_r)^2 + \Gamma_i^2} \quad (3.8)$$

After several lines of elementary algebra [see appendix c], we may write *Eqs.*(3.7) and (3.8) in forms which readily display the nature of the curves on  $\Gamma_r, \Gamma_i$  axes,

$$\left(\Gamma_r - \frac{r}{1+r}\right)^2 + \Gamma_i^2 = \left(\frac{1}{1+r}\right)^2 \quad (3.9)$$

$$(\Gamma_r - 1)^2 + \left(\Gamma_i - \frac{1}{x}\right)^2 = \left(\frac{1}{x}\right)^2 \quad (3.10)$$

And this is equation of circle with center  $(\Gamma_r, \Gamma_i) = (\frac{r}{1+r}, 0)$ , radius =  $\frac{1}{1+r}$  for the first equation .

And the center at  $(\Gamma_r, \Gamma_i) = (1, \frac{1}{x})$ , radius =  $\frac{1}{x}$  for the second equation

The first equation(3.9) describes a family of circles, where each circle is associated with a specific value of resistance  $r$ .

- The plot you usually see is the inside of the region bounded by the circle  $|\Gamma| = 1$

Outside this region there is reflection gain; in this outside region, the reflected signal is larger than the incident signal and this can only happen for  $r$  less than 0 (negative values of the real part of the load impedance). Thus the perimeter of the Smith chart as usually plotted is the  $r = 0$  circle, which is coincident with the  $|\Gamma| = 1$  circle.

For example, if  $r = 0$  the radius of this zero-resistance circle is seen to be unity, and it is centered at  $\Gamma_r = 0, \Gamma_i = 0$ , the origin.

This checks, for a pure reactance termination leads to a reflection coefficient of unity magnitude, and this for short circuit load .

On the other hand, if  $r = \infty$  , then  $z_L = \infty$  and we have  $\Gamma = 1 + j0$ .

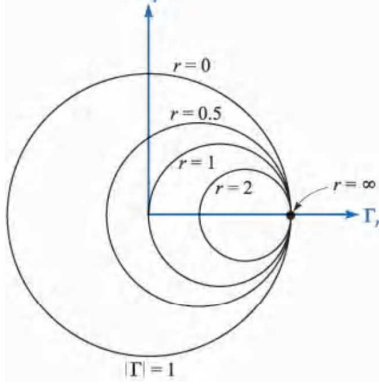


Figure 3.2: Constant- $r$  circles are shown on the  $\Gamma_r, \Gamma_i$  plane. The radius of any circle is  $1/(1+r)$ .

This circle described by Fig.(3.2) is centered at  $\Gamma_r = 1, \Gamma_i = 0$  and has zero radius. It is therefore the point  $\Gamma = 1 + j0$ , and this for open circuit load.

As another example, the circle for  $r = 1$  is centered at  $\Gamma_r = 0.5, \Gamma_i = 0$  and has a radius of 0.5. This circle is shown on Fig. (3.2), along with circles for  $r = 0.5$  and  $r = 2$ .

All circles one within the other are centered on the  $\Gamma_r$  axis and pass through the point  $\Gamma = 1 + j0$ .

Eq. (3.10) also represents a family of circles, but each of these circles is defined by a particular value of  $x$ , the constant reactance .

If  $x = \infty$  , then  $\Gamma = 1 + j0$  again. The circle described by Eq.(3.10) is centered at  $\Gamma = 1 + j0$  and has zero radius; it is therefore the point  $\Gamma = 1 + j0$ . If  $x = +1$ , then the circle is centered at  $\Gamma = 1 + j1$  and has unit radius. Only one-quarter of this circle lies within the boundary curve  $|\Gamma| = 1$ , as shown in Fig.(3.3) a similar quarter-circle appears below the  $\Gamma_r$ , axis for  $x = -1$ .

The portions of other circles for  $x = 0.5, -0.5, 2$ , and  $-2$  are also shown, for  $x$ -circles the outer part (+) for inductance and the lower part (-) for capacitance .

The circle representing  $x = 0$  is the  $\Gamma$  axis; this is also labeled on Fig.( 3.3) The two families of circles both appear on the Smith chart, as shown in Fig.(3.4)

It is now evident that if we are given  $Z_L$  we may divide by  $Z_o$  to obtain  $z_L$  locate the appropriate  $r$  and  $x$  circles (interpolating as necessary), and determine  $\Gamma$  by the intersection of the two circles. Since the chart does not have concentric circles

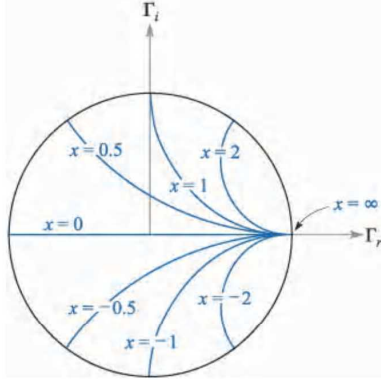


Figure 3.3: The portions of the circles of constant  $x$  lying within  $|\Gamma| = 1$  are shown on the  $\Gamma_r$ ,  $\Gamma_i$  axes. The radius of a given circle is  $1/|x|$ .

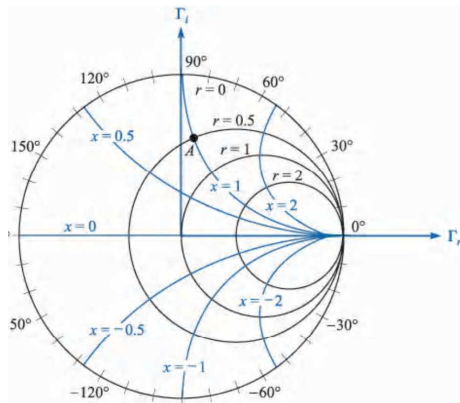


Figure 3.4: Constant  $r$  and  $x$  circles shown in it normalized load impedance  $z_L = 0.5 + j1$

showing the values of  $|\Gamma|$ , it is necessary to measure the radial distance from the origin to the intersection with dividers or compass and use an auxiliary scale to find  $\Gamma$ . The graduated line segment below the chart in Fig( 3.5) serves this purpose.

The angle of  $\Gamma$  is  $\phi$ , and it is the counter-clockwise angle from the  $\Gamma_x$ , axis. Again, radial lines showing the angle would clutter up the chart badly, so the angle is indicated on the circumference of the circle.

A straight line from the origin through the intersection may be extended to the perimeter of the chart.

As an example, if  $Z_L = 25 + j50 \Omega$  on a  $50\Omega$  line,  $z_L = 0.5 + j1$ , and point A on Fig.(3.4 ) shows the intersection of the  $r = 0.5$  and  $x = 1$  circles. The reflection coefficient is approximately 0.62 at an angle  $\phi$  of  $83^\circ$

The Smith chart represents both impedance and admittance plots. To use it as an admittance plot, turn it through 180 degrees about the center point. The directions towards the generator and towards the load remain in the same sense. The contours of constant resistance and constant reactance are now to be interpreted as constant (normalized) conductance  $g$ , and (normalized) susceptance  $s$  respectively. To see this property of the Smith chart we note first that the admittance  $y$  is the reciprocal of the impedance  $z_L$  (both being normalized). Thus inverting the equation above we see that.

$$y = g + js = \frac{1 - \Gamma}{1 + \Gamma} \quad (3.11)$$

and this is the same formula that we had above if we make the substitution  $\Gamma$  by  $-\Gamma$ . Of course, inverting the Smith Chart is the same as rotating it though 180 degrees or  $\pi$  radians, since  $-\Gamma = \Gamma(e^{j\pi})$ .

Admittance plots are useful for shunt connected elements; that is, for elements in parallel with the line and the load.[5]

$= 1$  are shown on the  $\Gamma_r$ ,  $\Gamma_i$  axes. The radius of a given circle is  $1/|x|$ .

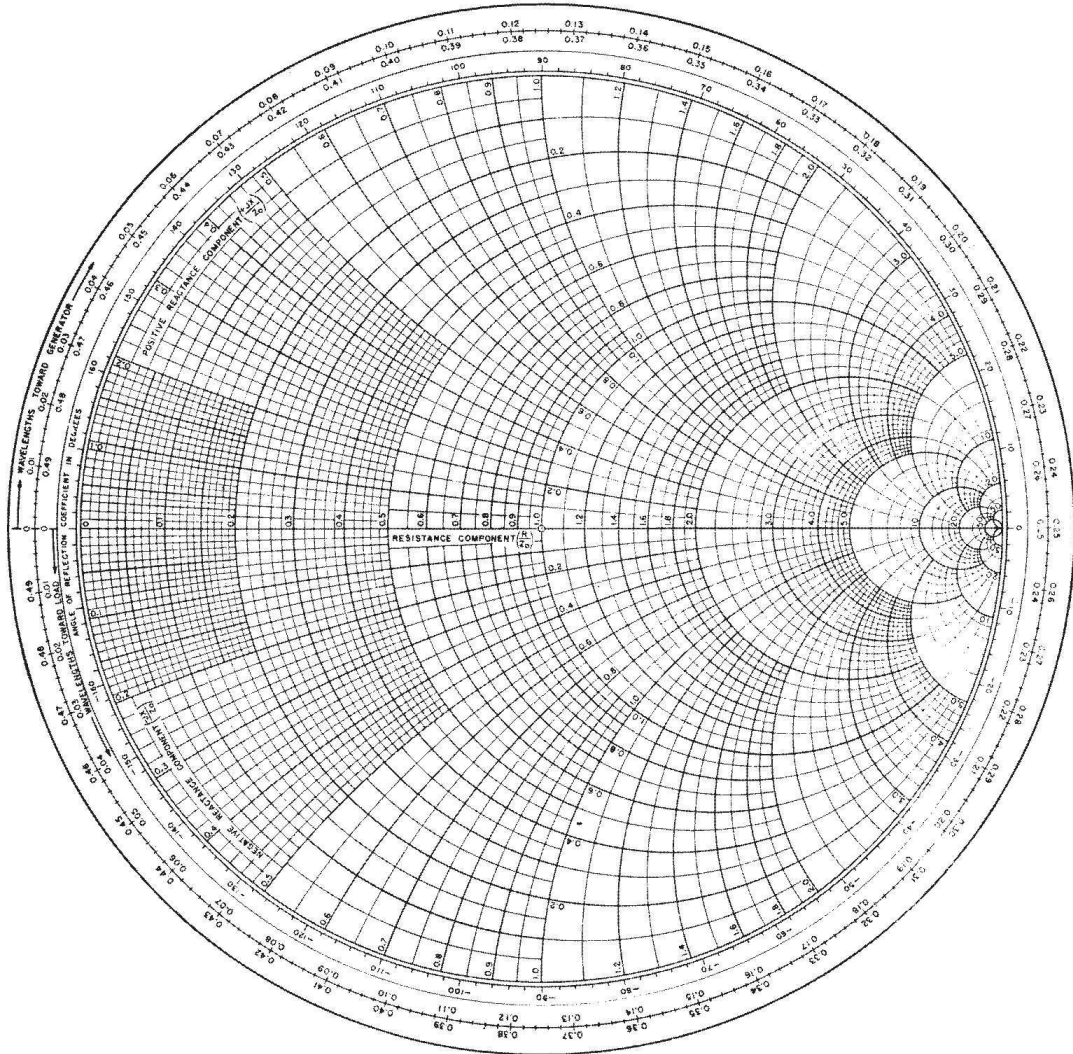


Figure 3.5: The Smith Chart contains the constant  $r$  circles and constant  $x$  circles, an auxiliary radial scale to determine  $|\Gamma|$ , and an angular scale on the circumference for measuring  $\phi$



# Chapter 4

## The Tunnel Diodes

### 4.1 INTRODUCTION

In this chapter we consider a device based on quantum-mechanical tunneling (the tunnel diodes).

In the classical sense, carriers having energy smaller than some potential barrier height are confined or stopped by the barrier completely. In quantum mechanics, the wave nature of carriers are considered.

As a result not only can the carriers have finite probability of [6,7] existence inside the barrier, they can leak through the barrier if the barrier width is thin enough. This leads to the concept of tunneling probability and tunneling current.

The tunneling process and devices based on this phenomenon have some interesting properties. First, the tunneling phenomenon is a majority-carrier effect, and the tunneling time of carriers through the potential energy barrier is not governed by the conventional transit time concept ( $\tau = W/v$ , where  $W$  is the barrier width and  $v$  is the carrier velocity), but rather by the quantum transition probability per unit time which is proportional to  $\exp[-2\langle k(O) \rangle W]$ , where  $\langle k(O) \rangle$  is the average value of momentum encountered in the tunneling path corresponding to an incident carrier with zero transverse momentum and energy equal to the Fermi energy. Reciprocity gives the tunneling time proportional to  $\exp[2\langle k(O) \rangle W]$ .

This tunneling time is very short, permitting the use of tunnel devices well into the millimeter-wave region. Secondly, since the tunneling probability depends on the available states of both the originating side and the receiving side, tunneling current is not monotonically dependent on the bias, and negative differential resistance can result.

The necessary conditions for tunneling are:

- Occupied energy states exist on the side from which the electron tunnels.
- Unoccupied energy states exist at the same energy level on the side to which the electron can tunnel.
- The tunneling potential barrier height is low and the barrier width is small enough that there is a finite tunneling probability.
- The momentum is conserved in the tunneling process.

## 4.2 The Tunnel Diode

The tunnel diode is  $p - n$  junction with heavily doped impurities and the doping levels from one hundred to several thousand times greater than  $p - n$  junction causing reduced depletion region, the tunnel diode is much more conductive at very small voltage, the tunnel diode is different from any known diode in that it has a negative resistance region.

In this region, an increase in terminal voltage results in a reduction in diode current, the depletion region is very thin so that many carriers can tunnel through, at low forward-bias potentials, the tunnel diode can be used in the manufacture of high speed devices as computers .

## 4.3 Resonant Tunneling Diode(RTD)

A resonant-tunneling diode requires band-edge discontinuity at the conduction band or valence band to form a quantum well [6]. The most-popular material combination used is GaAs/AlGaAs, followed by GaInAs/AlInAs. The middle quantum-well thickness is typically around 5 nm, and the barrier layers range from 1.5 to 5 nm. Symmetry of the barrier layers is not required, so their thicknesses can be different. The well and barrier layers are all undoped, and they are sandwiched between heavily doped, narrow-band gap materials, which usually are the same as the well layer. Layers are thin of undoped spacers  $\approx (1.5)$  nm *GaAs* adjacent to the barrier layers to ensure that dopants do not diffuse to the barrier layers.

## 4.4 Resonant Tunneling Diode - Nonlinear Transmission Lines (*RTD - NLTL*)

It is known that some active transmission lines have the property of shaping signal waveforms during their transmission, recently there has been great progress in the development of high -speed devices in electronics and optoelectronics.[8]

Furthermore, introducing nonlinear semiconductor devices into a distributes network results in nonlinear wave propagation can be used for short electrical pulse, generation periodic arrangement of resonant tunneling diode (*RTDs*) in a coplanar waveguide (*RTD - NLTL*) can be utilized as the basis of very interesting microwave signal generation circuits. The underlying characteristic of the *RTD* is a nonlinear N-shaped current -voltage relationship providing active behavior even at millimeter wave frequencies.

Fig.(4.1)is transmission line the circuit consists of  $N = 80$  element of (*RTD - NLTL*) it operates at millimeter wave oscillator. The equivalent circuit consists of as a shunt conductance -capacitance ( $J(V) - c$ ) circuit.

The active quantum well region of *RTD* and a series resistance -inductance ( $R - L$ ) circuit the nonlinear element is determined by the *RTD* current voltage relationship approximated by [8]

$$J(V) = BV(V - u_1)(V - u_2) \quad (4.1)$$

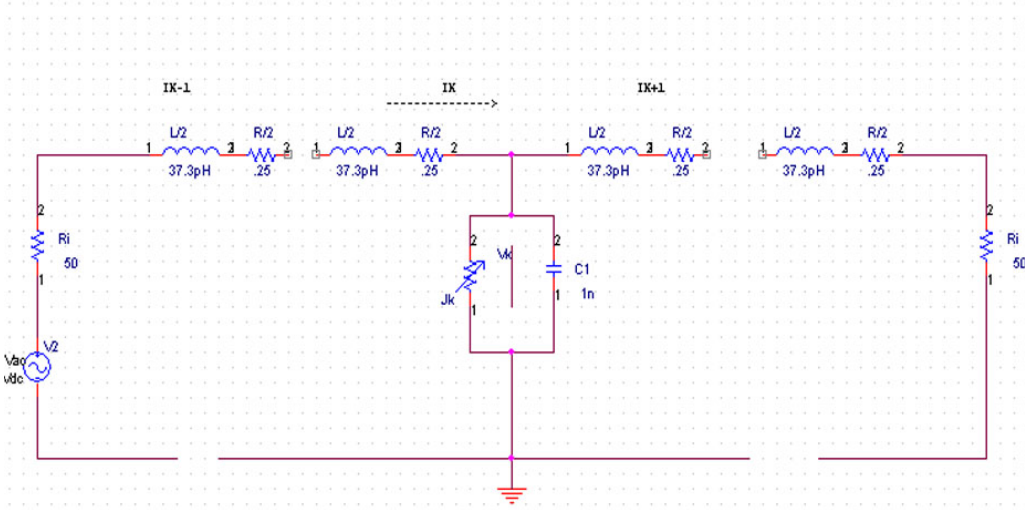


Figure 4.1: The equivalent circuit(RTD-NLTL)/Essimbi

where  $u_1, u_2$  are constants,  $(B)$  is a factor determined by the slop at  $V = 0$ ,  $J$  is current through  $(RTD)$

A set of equations which describes the voltage inside the line is given by Kirchhoff current law

$$C \frac{dV_k}{dt} = I_k - I_{k+1} - J_k \dots \dots \quad (4.2)$$

where  $k$  refired to number of element  $(RTD)$   
and Kirchhoff voltage law

$$L \frac{dI_k}{dt} = V_k - V_{k-1} - I_k R \dots \dots \quad (4.3)$$

The characteristic of  $(RTD)$  appears by different values of  $u_1, u_2$ .

The input is sinusoidal of frequency  $2GHz$  with amplitude  $0.5$  volt, the output may be rectangular or Square, in fact the signal is amplified in nonlinear differential resistance.

Figs.(4.2) show the  $V - I$  characteristics of  $(RTD - NLTL)$  for different values of  $u_1, u_2$  .

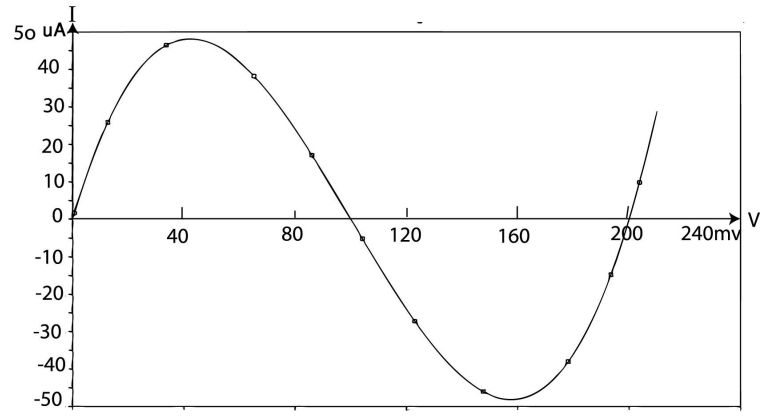
Fig.(4.2.a) represent  $V - I$  curve for  $u_1 = .1, u_2 = .2, s_1 = s_2$ ,

where  $s_1, s_2$  represented the positive and negative area for the  $J(V)$  curves .

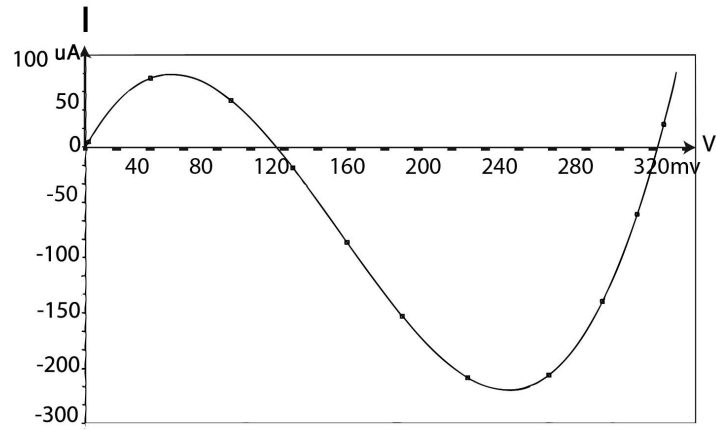
Fig.(4.2.b) is for  $u_1 = .1, u_2 = .3$ , in this case  $s_1 > s_2$ , and

Fig.(4.2.c) is for  $u_1 = .2, u_2 = .3$ , and in this case  $s_2 > s_1$ .

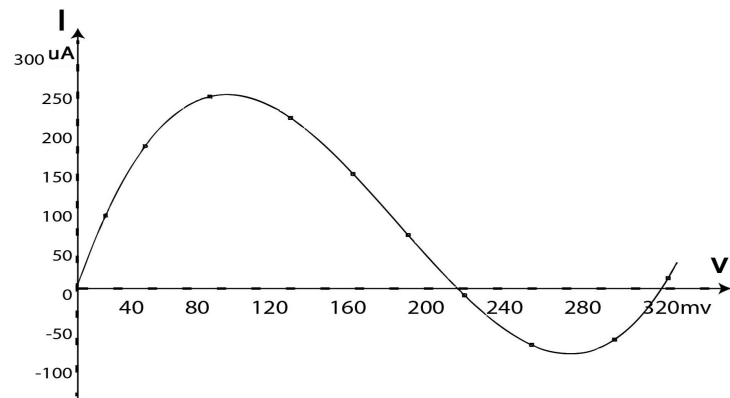
Using OrCad program to solve Eqs.(4.2, 4.3) for NLTL with number of units of RTD ( $N = 80$ ). The load resistance  $(R_l) = \infty$  Fig (4.3) shows the output curve at element 30 .The input signal is sinusoidal at frequency  $2GHz$  .The output signal are



((a))  $J(V) = 0.125 V (V-0.1)(V-0.2)$

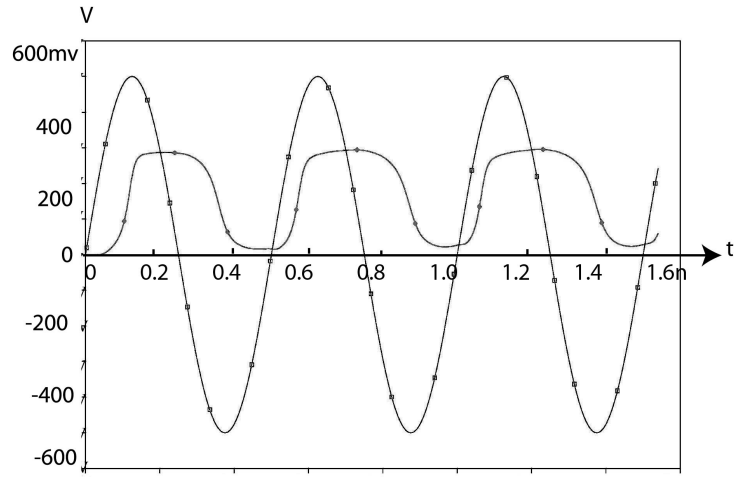


((b))  $J(V) = 0.125 V (V-0.1)(V-0.3)$

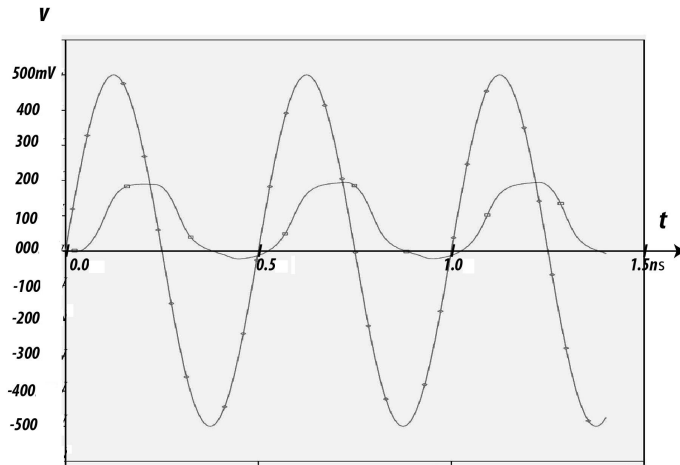


((c))  $J(V) = 0.125 V (V-0.2)(V-0.3)$

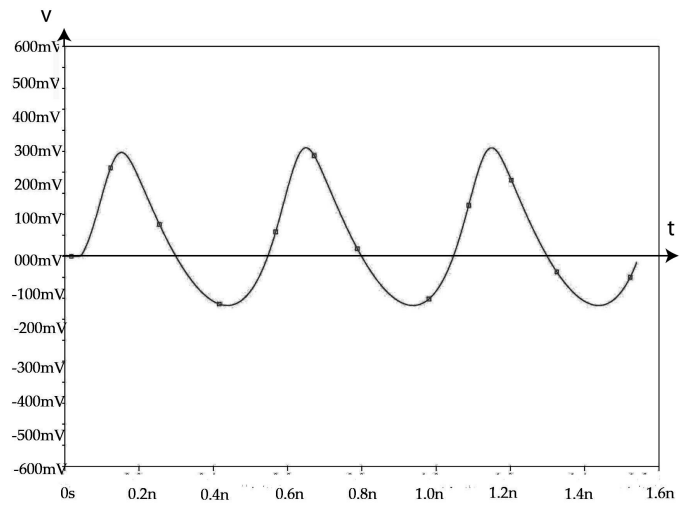
Figure 4.2: Characteristics of ( $RTD - NLTL$ )



((a))  $J(V) = 0.125 \text{ V } (V-0.1)(V-0.2)$



((b))  $J(V) = 0.125 \text{ V } (V-0.1)(V-0.3)$



((c))  $J(V) = 0.125 \text{ V } (V-0.2)(V-0.3)$

Figure 4.3: The output of sinusoidal pulse is square of  $K = 30, N = 80$

calculated for different values of  $u_1, u_2$ . Fig.(4.3.a) shows the output signal in volt versus time in nanosecond for  $u_1 = 0.1$  and  $u_2 = 0.2$ . Fig.(4.3.b) exhibit results for  $u_1 = 0.1$  and  $u_2 = 0.3$  Fig.(4.3.c) shows the output for  $u_1 = 0.2$  and  $u_2 = 0.3$  We see that Figs.(4.3.b) and (4.3.c) more stable than Fig.(4.3.a).And we can reshape the output signal as shown in Figs.(4.3) see Esmbi.[8]

## 4.5 The effect of a resistance on the output of sinusoidal pulse

I constricted a circuit as shown in Fig.(4.4.a) by OrCad and applied a resistance at the end of a circuit consists of  $N = 80$  element of  $RTD$ , and taken the output at element  $k = 30$  , we noticed that the amplitude of the output signal is smaller than thats in Fig.(4.3) with value of  $u_1 = 0.1$  and  $u_2 = 0.2$ , implies that exist a loss in the output as shown in Fig.(4.4.b)

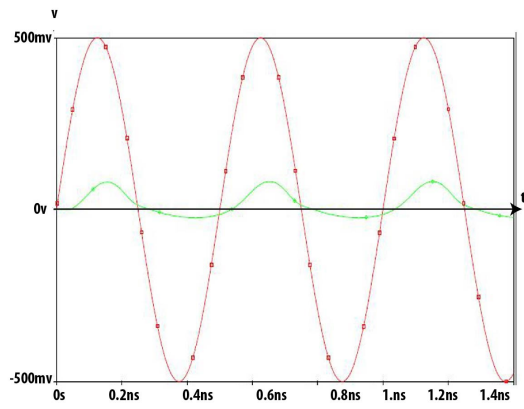
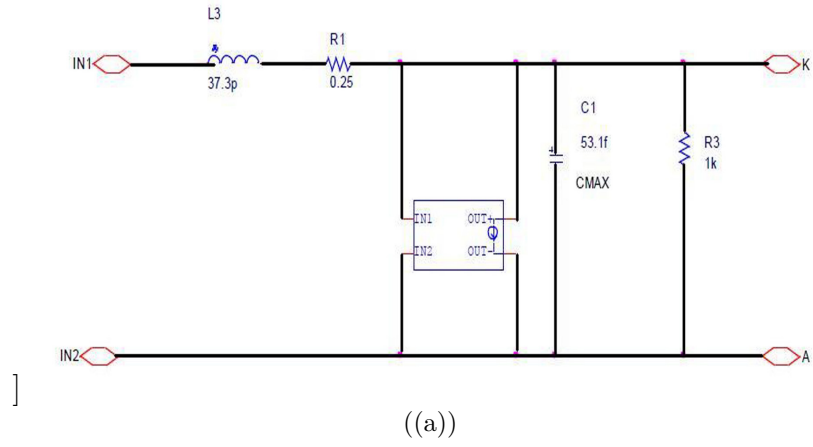


Figure 4.4: The effect of a resistance at  $J(V) = 0.125V(V - 0.1)(V - 0.2)$

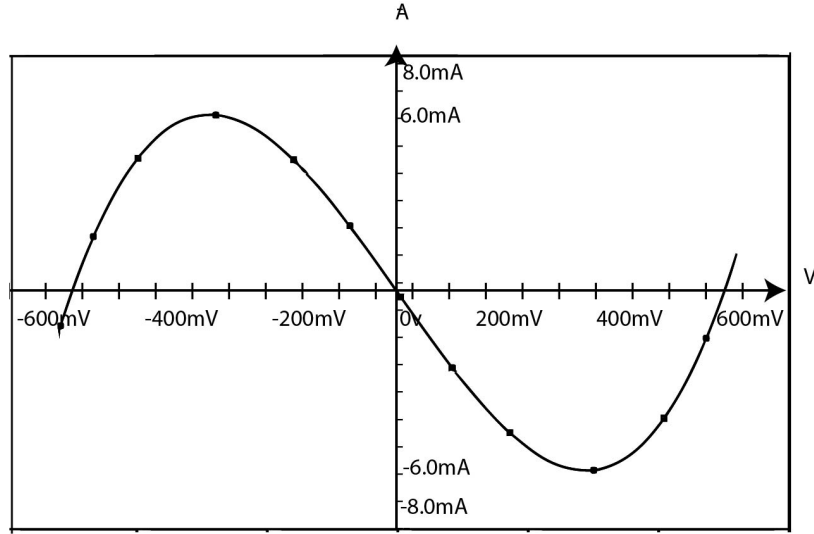


Figure 4.5:  $J(V) = 0.125V(V - 0.5)(V + 0.5)$

## 4.6 ( $RTD$ ) as an resonator:

If we take the relation, for one circuit

$$J(V) = BV(V - .5)(V + .5)$$

$$J(V) = BV(V^2 - .25)$$

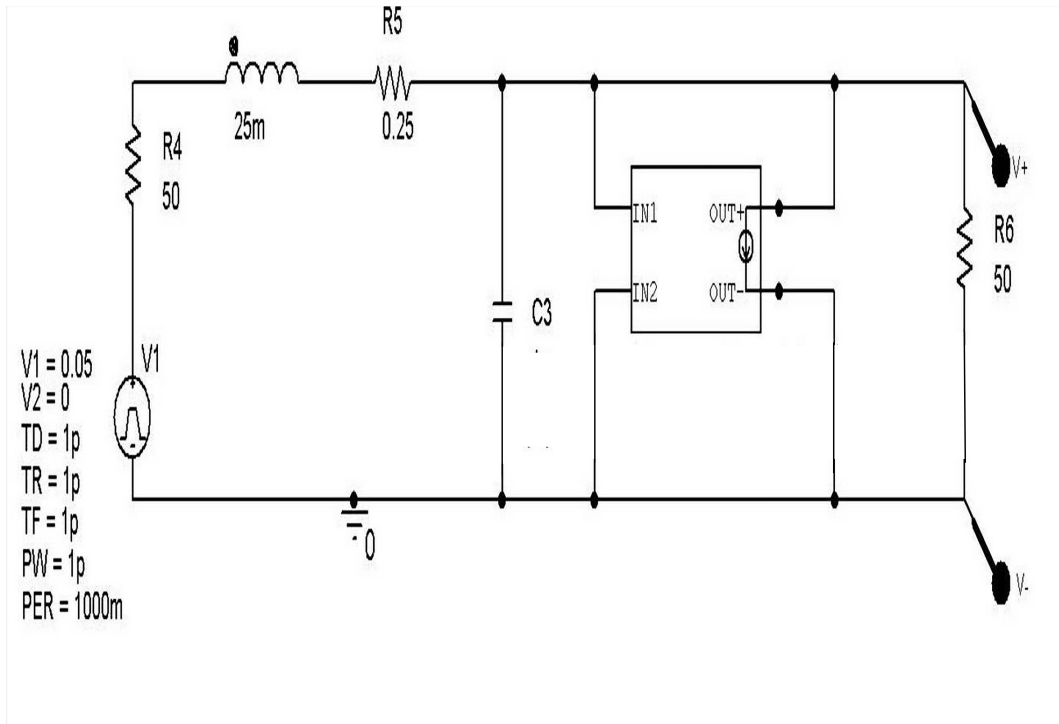
$$J(V) = B(V^3 - .25V)$$

Fig.(4.5 ) show ( $RTD - NDR$ )

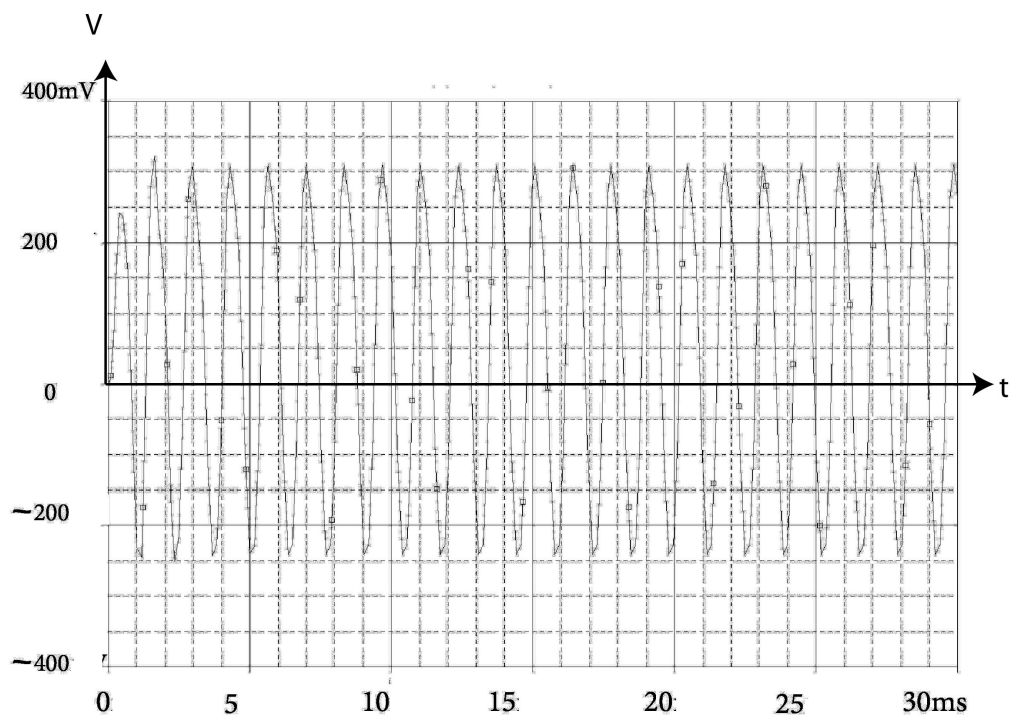
Using  $u_1 = u_2 = 0.5$  results is characteristic curve for  $RTD$  shows negative differential resistance ( $NDR$ ) which results an oscillator when applied to circuit with defined value of  $R_l$  .

The derived equations are solved using OrCad Fig(4.6). And this relation is similar to relation (4.4) in section (4.7) and this relation satisfy ( $VDP$ ) an oscillator and this nonlinear relation causes negative differential resistance .



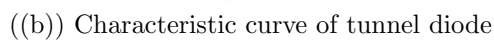


((a))



((b))

Figure 4.6: *RTD – Essimbi – Oscillator* :  $J(V) = 0.125V(V - 0.5)(V + 0.5)$



#### 4.7 Using Resonant-Tunneling Diode ( $RTD$ ) in an oscillator:

$$\dot{i}(t) = -a\nu(t) + b\nu^3(t) \quad (4.4)$$

We get:

$$i_D = i_s + i(t) \quad (4.5)$$

32

Kirchhoff's current rule states that the current  $i_L$  through inductor  $L$  is equal to sum of the currents through the resistor  $R$ , capacitor  $C$ , and the diode  $D$ .

$$i_L = i_R + i_C + i_D \quad (4.6)$$

$$i_L = i_R + i_C + i_s + i(t) \quad (4.7)$$

$$-i_L + i_R + i_C + i_s - a\nu(t) + b\nu^3(t) = 0 \quad (4.8)$$

The voltage drop across the resistor  $V_R$  and capacitor  $V_C$  are equal to voltage drop across the diode.

$$V_D = V_C = V_R = V_S + \nu(t) \quad (4.9)$$

where  $V_s$  is voltage source .The currents through resistor and capacitor are

$$i_R = V_R/R \quad (4.10)$$

$$i_C = C(\frac{dV_C}{dt}) \quad (4.11)$$

With all these relation entered and taking the time derivative of Eq. (4.8) we get:(4.12)

$$-(\frac{\partial}{\partial t}i_L) + \frac{1}{R}\frac{\partial}{\partial t}\nu(t) + C(\frac{\partial^2}{\partial t^2}\nu(t)) - a(\frac{\partial}{\partial t}\nu(t)) + 3b\nu^2(t)(\frac{\partial}{\partial t}\nu(t)) = 0 \quad (4.12)$$

From the definition of inductance, one has

$$\frac{di_L}{dt} = \frac{-V_L}{L} \quad (4.13)$$

By applying the kirchhoff's potential drop rule to outer loop of circuit we know that

$$V_L + V_D = V_S \quad (4.14)$$

$$V_L = V_S - V_D = \nu(t) \quad (4.15)$$

Substituting the induction definition and dividing Eq. (4.12) by C yields:Eq.(4.16)

$$(\frac{1}{CR} - \frac{a}{C} + \frac{3b\nu^2(t)}{C})(\frac{\partial}{\partial t}\nu(t)) + \omega^2\nu(t) + \frac{\partial^2}{\partial t^2}\nu(t) = 0 \quad (4.16)$$

where  $\omega = \frac{1}{\sqrt{LC}}$  is the frequency

By introducing a dimensionless time variable  $\tau$  defined by  $t = \tau/\omega$  and a dimensionless voltage variable defined by

$$\nu(t) = X(\tau) \frac{\sqrt{a - 1/R}}{\sqrt{3b}} = X(\tau)t_r \quad (4.17)$$

Eq.(4.16) reduces to Van Der Pol equation.

$$\left(\frac{\partial^2}{\partial \tau^2} X(\tau)\right) - \mu(1 - X^2(\tau))\left(\frac{\partial}{\partial \tau} X(\tau) + X(\tau)\right) = 0 \quad (4.18)$$

where  $\mu$  positive dimensionless parameter related to the circuit parameters with value

$$\mu = \frac{aR - 1}{\omega CR} \quad (4.19)$$

$\omega$  controls how much voltage is injected into the system.  $\mu$  controls the way in which voltage flows through the system.[11,12,13]

The Van der Pol oscillator is a classical example of self oscillator system and is now considered as very useful mathematical model that can be used in much more complicated and modified systems. The van der pol equation has played an important role, particularly because it displays the so-called limit cycle, a phenomenon which does not occur in linear problems. A limit cycle, corresponds to a periodic motion which the system approaches no matter what the initial conditions are. All electronic oscillator circuits display limit cycles as do many acoustical and mechanical systems, the heart, a mechanical pump being a prime example. the Van der Pol (vdP) equation, for a particular electronic oscillator circuit, can be regarded as a simple harmonic equation.

The equation of simple pendulum is written as

$$\ddot{\Theta} + \omega_0^2 \sin \Theta = 0 \quad (4.20)$$

$$\omega_o = \sqrt{\frac{g}{l}} \quad (4.21)$$

Where  $\Theta$  is angular accelerations and this is harmonica motion and the equation of van Der Pol is similar to this equation

The van der Pol equation, in what is now considered to be standard form, is given by

$$\ddot{x} - \mu(1 - x^2)\dot{x} + x = 0. \quad (4.22)$$

We see that it is an oscillator with a linear spring force and a nonlinear damping force. The damping force varies in an interesting way. For  $\mu < 1$ , the damping is actually negative and hence produces an amplification of the motion. For  $\mu > 1$ , there is true damping and the motion decays. These observations suggest the possibility of an oscillation, in which the system starts at small  $x$ , is driven to large  $x$  by the amplification, and is then damped back to small  $x$ . We can convert the equation to the following system:

$$\dot{x} = y, \dot{y} = -x + \mu(1 - x^2)y$$

## 4.8 Solving the Van Der Pol equation with Runge kutta algorithm

$$\begin{aligned}k_1 &= f(t_n, x_n, y_n) \\I_1 &= g(t_n, x_n, y_n) \\K_2 &= f(t_n + \frac{1}{2}h, x_n + \frac{1}{2}h * k_1, y_n + \frac{1}{2}h * I_1) \\I_2 &= g(t_n + \frac{1}{2}h, x_n + \frac{1}{2}h * k_1, y_n + \frac{1}{2}h * I_1) \\K_3 &= f(t_n + \frac{1}{2}h, x_n + \frac{1}{2}h * k_2, y_n + \frac{1}{2}h * I_2) \\I_3 &= g(t_n + \frac{1}{2}h, x_n + \frac{1}{2}h * k_2, y_n + \frac{1}{2}h * I_2) \\K_4 &= f(t_n + h, x_n + h * k_3, y_n + h * I_3) \\I_4 &= g(t_n + h, x_n + h * k_3, y_n + h * I_3) \\k &= \frac{1}{6}(k_1 + 2k_2, +2k_3, +k_4) \\I &= \frac{1}{6}(I_1 + 2I_2, +2I_3, +I_4) \\x_{n+1} &= x_n + h * k \\y_{n+1} &= y_n + h * I \\t_{n+1} &= t_n + h\end{aligned}$$

where  $f(t_n, x_n, y_n) = y$

$$g(t_n, x_n, y_n) = -x + \mu(1 - x^2)y$$

This is Runge Kutta algorithm for second ordinary differential equation and can always be rewritten as two equivalent first- order equations .

Runge Kutta method is used for numerical solution of ordinary differential equation of first-order at accurate approximation.

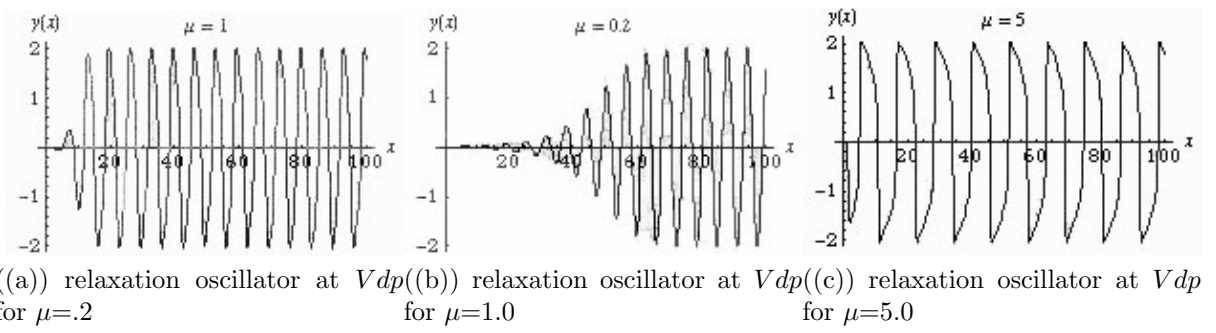


Figure 4.8: Relaxation oscillator at  $Vdp$

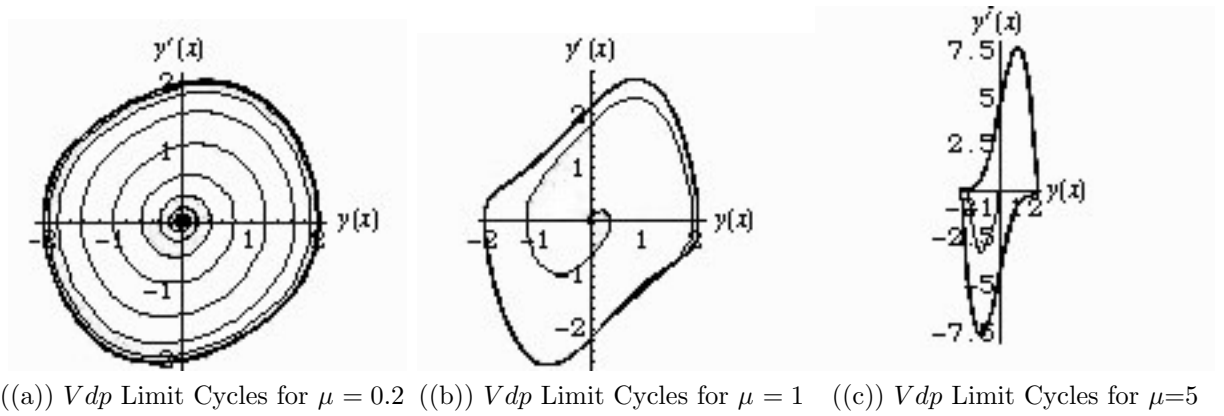


Figure 4.9:  $Vdp$  Limit Cycles

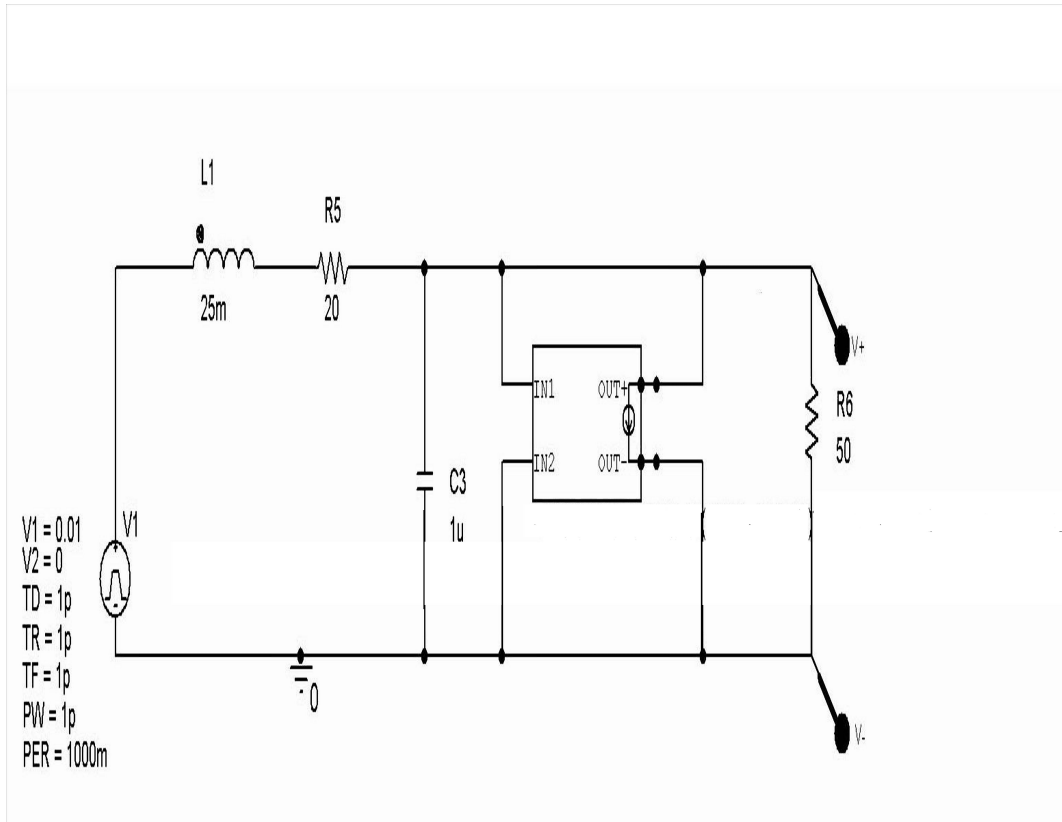
## 4.9 Using Mathematica Programming

I have taken a ready a program in mathematica containing (Runge Kutta package)  
 [see appendix D]

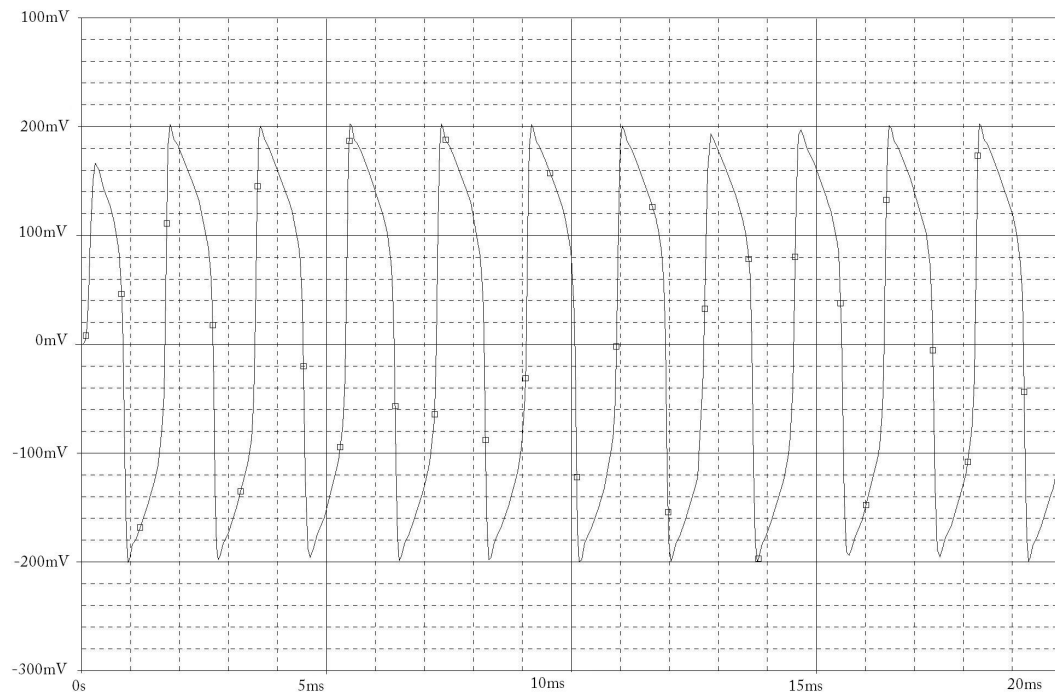
Fig.(4.8)is plotted by numerically solving the Van der Pol equation to different amount of  $\mu$ ,implies thats this is relation between voltage ( $y(x)$ )and time ( $x$ )in the graph .Fig.(4.9) is relation between voltage ( $y(x)$ ) and  $\dot{y}(x) = \dot{v}$  for different values of  $\mu$  .

If we have a load we get at the output an oscillator Fig.(4.10.b)

Fig.(4.10.b)is plotted by Orcad simulation showing eq.(4.18) is VDP oscillator



((a)) 22



((b)) 33

Figure 4.10:  $Vdp - RTD$  oscillator:  $i(t) = 0.050v(t) + 1.0v^3(t)$



# Chapter 5

## Schottky Diodes

### 5.1 Metal-Semiconductor (MS) Junctions

Schottky Diodes consist of a metal contacting a piece of semiconductor in one side. An ideal ohmic contact, a contact such that no potential exists between the metal and the semiconductor(doesn't rectify current), is made to the other side of the semiconductor, the structure is shown in Fig.(5.1)

Schottky diode is an important power device and used as output rectifiers in switching-mode power supplies and in other high-speed power switching applications, such as motor drives, switching of communication device, industry automation and electronic automation and so on.[6]

The use of a Schottky diode generally allows integrated circuits to have greater speed because it is a majority carrier device.

The Schottky diodes have lower turn-on voltages and because of the lower barrier height of the rectifying metal-to-semiconductor junction it have faster switching speeds because they are primarily majority carrier devices.

### 5.2 Energy Band Diagram And Built-In Potential

The barrier between the metal and the semiconductor can be identified on an energy band diagram. To construct such diagram we first consider some concepts .[7,14,15,16]

- Fermi energy( $E_f$ ) is the energy of highest occupied state of electrons .

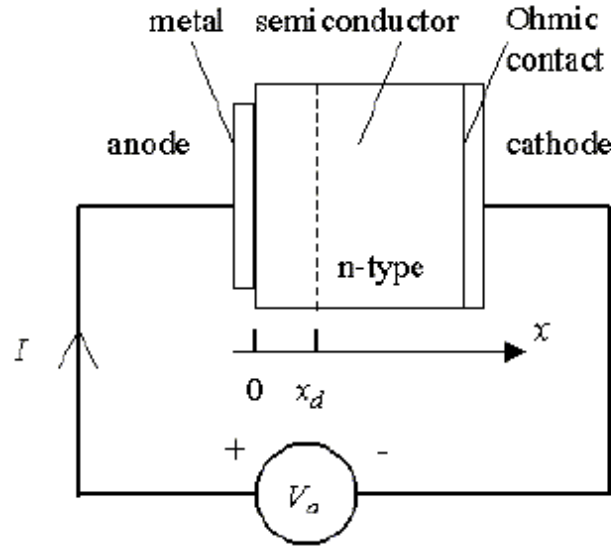


Figure 5.1: A metal-semiconductor junction(Schottky Diode)

- Free electron energy ( $E_0$ ) is the energy needed by an electron to escape from the surface of the material .
- Work function( $\phi$ ) is the energy difference between the free electron energy and the average electron energy(or Fermi)level.

$$q\phi = E_0 - E_f$$

since Fermi energy varies with the type of amount of doping in the semiconductor so does the work function.

- The electron affinity ( $\chi$ ) is defined as the difference between the bottom of the conduction band ( $E_c$ ) and free electron energy ( $E_0$ ).

$$q\chi = E_0 - E_c$$

since  $E_c = E_f$  in the metal, then the electron affinity is the same as its work function .

the energy band diagram of the metal as shown in Fig. (5.2 .a) and (5.2. b).

The barrier height,  $\phi_B$ , is defined as the potential difference between the Fermi energy of the metal and the band edge where the majority carriers reside. For an n-type semiconductor the barrier height is obtained from:

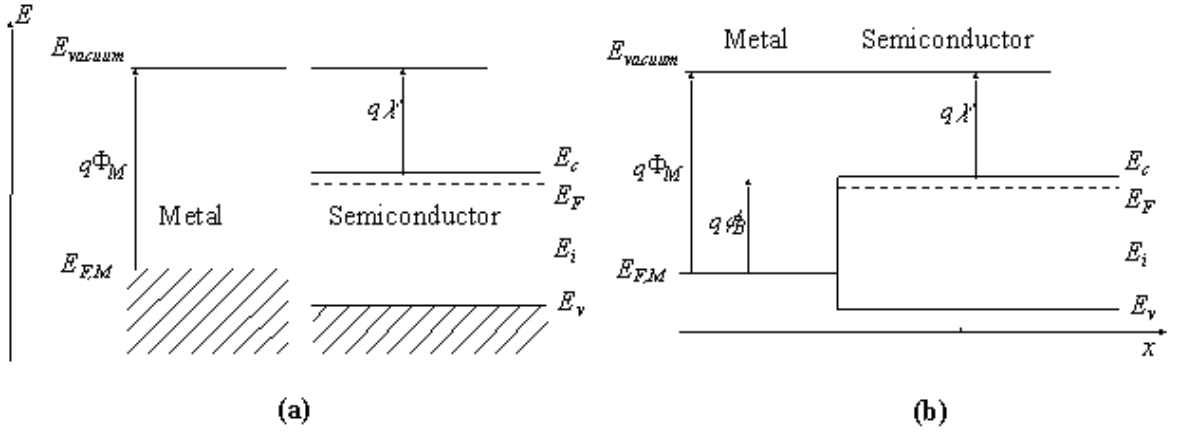


Figure 5.2: Energy band diagram of the metal and the semiconductor before contact (a) and after contact is made (b)

$$\phi_B = \Phi_M - q\chi \quad (5.1)$$

where  $\Phi_M$  is the work function of the metal and  $\chi$  is the electron affinity.

For p-type material, the barrier height is given by the difference between the valence band edge of the semiconductor and the Fermi energy in the metal:

$$\phi_B = E_g + q\chi - \Phi_M \quad (5.2)$$

where  $E_g$  is the energy gap

$$E_g = E_v - E_c$$

The built-in potential,  $\phi_I$ , as the difference between the Fermi energy of the metal and that of the semiconductor.

$$\phi_I = \Phi_M - q\chi - (E_c - E_{F,n}) \quad (5.3)$$

$$\phi_I = q\chi + (E_c - E_{F,p}) - \Phi_M \quad (5.4)$$

Eq.(5.3) for  $n$ -type, Eq.(5.4) for  $p$ -type,  $E_{F,n}$  is (fermi energy for  $n$ -type).

The barrier heights often differ experimentally from that ones calculated using Eqs.(5.1) or (5.2).

This is due that the ideal metal-semiconductor theory assumes that both materials are infinitely pure, that there is no interaction between the two materials

nor is there an interfacial layer. Chemical reactions between the metal and the semiconductor alter the barrier height[6,18] .

### 5.3 Depletion layer in M - S diode

It is of interest to note that in thermal equilibrium, i.e. with no external voltage applied, there is a region in the semiconductor close to the junction , which is depleted of mobile carriers. We call this the depletion region. The potential across the semiconductor equals the built-in potential,  $\phi_I$ . [6,18]

We now apply the full depletion approximation to an  $M - S$  junction containing an n-type semiconductor.

The semiconductor is fully depleted over a distance  $x_d$ , called the depletion region. While this assumption does not provide an accurate charge distribution, it does provide very reasonable approximate expressions for the electric field and potential throughout the semiconductor.

We define the depletion region to be between the metal-semiconductor interface ( $x = 0$ ) and the edge of the depletion region ( $x = x_d$ ).

As the semiconductor is depleted of mobile carriers within the depletion region, the charge density in that region is due to the ionized donors. Outside the depletion region, the semiconductor is assumed neutral. This yields the following expressions for the charge density,  $\rho$ :

$$\rho(x) = qN_d \dots\dots\dots 0 \leq x \leq x_d \dots\dots\dots \rho(x) = 0 \dots\dots\dots x_d < x \quad (5.5)$$

where  $q$  is electron charge and  $N_d$  is donor density.

where we assumed full ionization so that the ionized donor density equals the donor density,  $N_d$ . The charge in the semiconductor is exactly balanced by the charge in the metal,  $Q_M$ , so that no electric field exists except around the metal-semiconductor interface.

Using Gauss's law we obtain the electric field as a function of position,

Gauss's law in one direction

$$\frac{dE(x)}{dx} = \frac{\rho}{\epsilon_s} \quad (5.6)$$

where  $\epsilon_s$  is the dielectric constant of the semiconductor. taking the integral of Eq.(5.6)

$$\oint E \cdot dA = \frac{Q}{\epsilon_s} \quad (5.7)$$

$$E(x) = -\frac{qN_d}{\epsilon_s}(x_d - x) \dots \dots 0 \leq x \leq x_d \quad (5.8)$$

$$E(x) = 0 \dots \dots x_d < x \quad (5.9)$$

We also assumed that the electric field is zero outside the depletion region, since a non-zero field would cause the mobile carriers to redistribute until there is no field.

The depletion region does not contain mobile carriers so that there can be an electric field [6].

The largest (absolute) value of the electric field is obtained at the interface and is given by:

$$E(x=0) = -\frac{qN_dx_d}{\epsilon_s} = -\frac{Q_d}{\epsilon_s} \quad (5.10)$$

where the electric field was also related to the total charge (per unit area),  $Q_d$ , in the depletion layer. Since the electric field is minus the gradient of the potential,

$$E = -\nabla\phi(x) \quad (5.11)$$

taking the integrall

$$\phi(x) = -\int E dx \quad (5.12)$$

From  $x = x_d - x$  to  $x = x_d$

$$\phi(x) = \frac{qN_d}{\epsilon_s} \int x dx \quad (5.13)$$

$$\phi(x) = 0 \dots x \leq 0$$

$$\phi(x) = \frac{qN_d}{2\epsilon_s} [x_d^2 - (x_d - x)^2] \dots 0 < x < x_d \quad (5.14)$$

$$\phi(x) = \frac{qN_dx_d^2}{2\epsilon_s} \dots x_d \leq x \quad (5.15)$$

This potential due the semiconductor and the potential across the metal can be neglected. Since the density of free carriers is very high in a metal, the thickness of the charge layer in the metal is very thin.

The total potential difference across the semiconductor equals the built-in potential,  $\phi_i$ , in thermal equilibrium ( without external voltage ) and is further reduced (when positive voltage is applied ( $V_a$ ) to metal (forward) or increased when negative voltage is applied to the metal (reverse) and equal  $\phi_i - V_a$

This boundary condition provides the following relation between the semiconductor potential at the surface, the applied voltage and the depletion layer width

$$\phi_i - V_a = -\phi(x = x_d) = \frac{qN_dx_d^2}{2\epsilon_s} \quad (5.16)$$

Solving this expression for the depletion layer width,  $x_d$ , yields:

$$x_d = \sqrt{\frac{2\epsilon_s(\phi_i - V_a)}{qN_d}} \quad (5.17)$$

## 5.4 Junction capacitance of M-S diode

In addition, we can obtain the capacitance as a function of the applied voltage by taking the derivative of the charge with respect to the applied voltage yielding:[6,18]

$$C_j = \left| \frac{dQ_d}{dV_a} \right| \quad (5.18)$$

where  $Q_d = qN_dx_d$  ,  $x_d = \sqrt{\frac{2\epsilon_s(\phi_i - V_a)}{qN_d}}$  and  $C_j = qN_d \frac{dx_d}{dv_a}$

$$C_j = \frac{\epsilon_s}{x_d}$$

We can see that capacitance is not constant because the depletion layer width,  $x_d$ , varies with the applied voltage.

## 5.5 Diffusion Current

The diffusion theory assumes that the driving force is distributed over the length of the depletion layer.

This analysis assumes that the depletion layer is large compared to the mean free path, so that the concepts of drift and diffusion are valid. The resulting current density equals.[6,18]

$$J_n = \frac{q^2 D_n N_c}{V_t} \sqrt{\frac{2q(\phi_i - V_a)N_d}{\epsilon_s}} \exp\left(-\frac{\phi_B}{V_t} \left[\exp\left(\frac{V_a}{V_t}\right) - 1\right]\right) \quad (5.19)$$

Where  $D_n = \frac{KT\mu_n}{q}$ ,  $D_n$  is electron diffusion constant,  $\mu_n$  is mobility of an electron and  $v = \mu_n E$ ,  $v$  is the velocity,  $E$  is the electric field,  $N_o$  is the density of available carriers.

We can get this result from some relations .

$$J_n = (\mu_n n E + D_n \frac{dn}{dx}) \quad (5.20)$$

$E = \frac{-d\phi}{dx}$ , multiply by  $e^{\frac{-\phi}{v_t}}$  both sides and considering boundary conditions we can get the above relation and we are not concern for all details.

Where  $v_t$  is the thermal voltage at saturation  $v_t = \frac{KT}{q}$ ,  $K$  is Boltzman constant and  $T$  is absolute temperature .

The current therefore depends exponentially on the applied voltage,  $V_a$ , and the barrier height,  $\phi_B$ . The pre factor can more easily be understood if one rewrites it as a function of the electric field at the metal-semiconductor interface,  $E_{max}$  such that

$$E_{max} = \sqrt{\frac{2q(\phi_i - V_a)N_d}{\epsilon_s}} \quad (5.21)$$

yielding:

$$J_n = q\mu_n E_{max} N_c \exp\left(-\frac{\phi_B}{V_t}\right) \left[\exp\left(\frac{V_a}{V_t}\right) - 1\right] \quad (5.22)$$

## 5.6 Thermionic Emission

The thermionic emission theory assumes that electrons, which have an energy larger than the top of the barrier, will cross the barrier provided they move towards the barrier, the actual shape of the barrier is ignored and the current can be expressed as.

$$J_{MS} = A^* T^2 \exp\left(-\frac{\phi_B}{V_t}\right) \left[\exp\left(\frac{V_a}{V_t}\right) - 1\right] \quad (5.23)$$

where

$$A^* = \frac{4\pi q m^* k^2}{h^3} \quad (5.24)$$

$A^*$  is the Richardson constant and  $\phi_B$  is the Schottky barrier height,  $T$  is the temperature .

## 5.7 Tunneling

Quantum-mechanical tunneling through the barrier takes into account the wave-nature of the electrons, allowing them to penetrate through thin barriers.[6,18]

The tunneling current is obtained from the product of the carrier charge, velocity and density. The velocity equals the Richardson velocity, the velocity with which on average the carriers approach the barrier. The carrier density equals the density of available electrons,  $n$ , multiplied with the tunneling probability,  $\Theta$ , yielding:

$$J_n = q v_R n \Theta \quad (5.25)$$

where  $v_R$  is

$$v_R = \sqrt{\frac{kT}{2\pi m}}$$

Where the tunneling probability is obtained from:

$$\Theta = \exp\left(-\frac{4}{3} \frac{\sqrt{2q m^*}}{h} \frac{\phi_B^{3/2}}{E}\right) \quad (5.26)$$

and the electric field equals

$E = \phi_B / L$  Where  $L$  is the length of the well of the barrier. The tunneling current therefore depends exponentially on the barrier height,  $\phi_B$ , to the 3/2 power



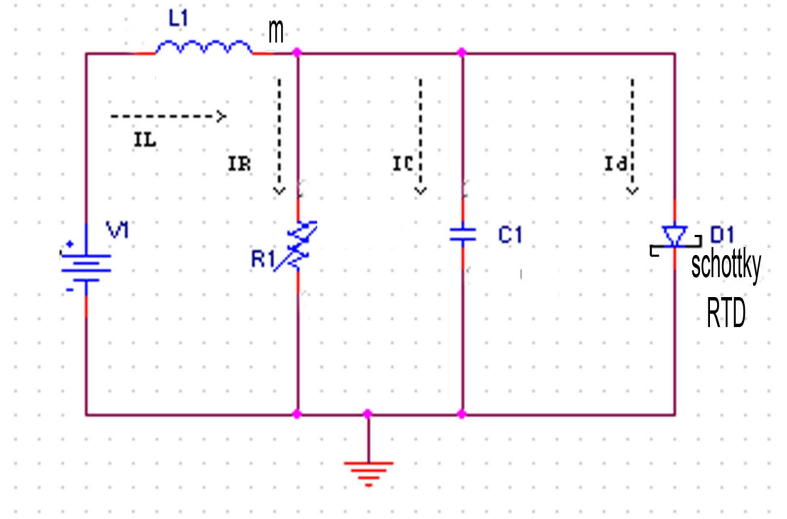


Figure 5.3: Schottky diode circuit

this result from Schrodinger equation and we are not concern with it her, In a given junction, a combination of all three mechanisms could exist.

## 5.8 Dispersion, Nonlinearity, and Dissipation in Schottky diodes

Schottky diodes provide nonlinearity due to the voltage dependent capacitance, dispersion due to the periodicity, and dissipation due to the finite conductivity of the CPW conductor and series resistance of the diodes.

## 5.9 Schottky Diode in An Oscillator

This work is aimed to study generation signals from an oscillator using Resonant-Schottky Diode.

We can use Schottky diode as a tunnel diode in special case in manufactures when the ratio of impurities is very high and the metal layer is very thin. we may write[14]

$$i(t) = I_{sat} \exp^{\delta v_t} (1 + \delta(v - v_t) + \frac{\delta}{2}(v - v_t)^2) \quad (5.27)$$

where  $I_{sat}$  is saturation current

$$I_{sat} = A^{**}T^2W_j \exp(\frac{q\phi_b}{KT}) \quad (5.28)$$

Where  $A^{**}$  modified Richardson constant, T is the absolute temperature and K is Boltzman constant .

$$A^{**} = 4.4Acm^{-2}k^{-2} \text{ for } GaAs$$

$W_j$  is junction area,  $v_t$  is threshold voltage .

$$\phi_b \text{ barrier hight in volts} = .7325v$$

$$\delta = \alpha * v \dots \dots \dots \text{where} \dots \dots \alpha = \frac{q}{\eta Kt} \quad (5.29)$$

$\eta$  is ideality factor  $1.2 > \eta > 1$

$$i_d = i_s + i(t) \quad (5.30)$$

Applying kirchhoff current law at node (m), see Fig.(5.3)

$$i_l = i_R + i_C + i_d \quad (5.31)$$

Substituting from Eqs.(5.27),(5.30) in(5.31) we get

$$-i_l + i_R + i_C + i_s + I_{sat} \exp^{\alpha v_t} (1 + \alpha * v(v - v_t) + \frac{\alpha * v}{2}(v - v_t)^2) = 0 \quad (5.32)$$

Voltage drop across R and C and d are equal

$$v_d = v_C = v_R = v_s + v(t) \quad (5.33)$$

And we have

$$i_R = \frac{v_R}{R}, i_C = C(\frac{dv_C}{dt}) \quad (5.34)$$

Substituted those relations in equation (5.32)

and take derivative with respect to time we have

$$-\frac{\partial i_l}{\partial t} + \frac{1}{R} \frac{\partial v(t)}{\partial t} + C \frac{\partial^2 v(t)}{\partial t^2} + \frac{\partial}{\partial t} \{ I_{sat} \exp^{\alpha v v_t} (1 + \alpha * v(v - v_t) + \frac{\alpha * v}{2} (v - v_t)^2) \} = 0 \quad (5.35)$$

take

$$\frac{\partial}{\partial t} \{ I_{sat} \exp^{\alpha v v_t} (1 + \alpha * v(v - v_t) + \frac{\alpha * v}{2} (v - v_t)^2) \} \quad (5.36)$$

$$= I_{sat} \{ (\frac{\partial}{\partial t} \exp^{\alpha v v_t}) (1 + \alpha * v(v - v_t) + \frac{\alpha * v}{2} (v - v_t)^2) \}$$

$$+ \exp^{\alpha v v_t} \frac{\partial}{\partial t} (1 + \alpha * v(v - v_t) + \frac{\alpha * v}{2} (v - v_t)^2) \} \quad (5.37)$$

$$= I_{sat} \{ \alpha v_t \exp^{\alpha v v_t} \{ \frac{\partial v}{\partial t} \} (1 + \alpha * v(v - v_t) + \frac{\alpha * v}{2} (v - v_t)^2) + \exp^{\alpha v v_t} (\alpha \{ \frac{\partial v}{\partial t} \} (v - v_t)) \}$$

$$+ \alpha v \frac{\partial v}{\partial t} + \frac{\alpha}{2} \{ \frac{\partial v}{\partial t} \} (v - v_t)^2 + \alpha v (v - v_t) \{ \frac{\partial v}{\partial t} \} \quad (5.38)$$

$$= I_{sat} \alpha \exp^{\alpha v v_t} \frac{\partial v}{\partial t} \{ v_t (1 + \alpha * v(v - v_t) + \frac{\alpha * v}{2} (v - v_t)^2) +$$

$$(v - v_t) + v + \frac{1}{2} (v - v_t)^2 + v(v - v_t) \} \quad (5.39)$$

$$\begin{aligned}
&= I_{sat} \alpha \exp^{\alpha v_t} \frac{\partial v}{\partial t} \left\{ v(-\alpha v_t^2 + \frac{\alpha}{2} v_t^3 - 2v_t + 2) \right. \\
&\quad \left. + v^2(\alpha v_t - \alpha v_t^2 + \frac{3}{2}) + v^3 \frac{\alpha}{2} v_t + \frac{1}{2} v_t^2 \right\} \tag{5.40}
\end{aligned}$$

$$= I_{sat} \exp^{\alpha v_t} \frac{\partial v}{\partial t} \{a_1 v + a_2 v^2 + a_3 v^3 + a_4\} \tag{5.41}$$

Where

$$a_1 = \alpha(-\alpha v_t^2 + \frac{\alpha}{2} v_t^3 - 2v_t + 2) \tag{5.42}$$

$$a_2 = \alpha(\alpha v_t - \alpha v_t^2 + \frac{3}{2})$$

$$a_3 = \frac{\alpha^2 v_t}{2}$$

$$a_4 = \frac{\alpha v_t^2}{2}$$

From the definition of inductance we have

$$\frac{\partial i_l}{\partial t} = \frac{v_l}{l} \tag{5.43}$$

$$\frac{v_l}{l} + \frac{1}{R} \frac{\partial v(t)}{\partial t} + C \frac{\partial^2 v(t)}{\partial t^2} + I_{sat} \exp^{\alpha v_t} \{a_1 v + a_2 v^2 + a_3 v^3 + a_4\} \frac{\partial v}{\partial t} = 0 \tag{5.44}$$

Divided by C

$$\frac{v_l}{Cl} + \frac{1}{CR} \frac{\partial v(t)}{\partial t} + \frac{\partial^2 v(t)}{\partial t^2} + \frac{1}{C} I_{sat} \exp^{\alpha v v_t} \{a_1 v + a_2 v^2 + a_3 v^3 + a_4\} \frac{\partial v}{\partial t} = 0 \quad (5.45)$$

$$\frac{\partial^2 v(t)}{\partial t^2} + \left\{ \frac{1}{CR} + \frac{1}{C} I_{sat} \exp^{\alpha v v_t} \{a_1 v + a_2 v^2 + a_3 v^3 + a_4\} \right\} \frac{\partial v(t)}{\partial t} + \omega^2 v(t) = 0 \quad (5.46)$$

Where

$$\omega = \frac{1}{\sqrt{IC}} \quad (5.47)$$

Eq.(5.46)for Schottky is a general formula for an oscillator and is similar to Eq.(4.18) for VDP Oscillator and we can get the same results .

# Conclusion

This work is aimed to study generation signals from an oscillator using Schottky resonant-tunneling diode by deducing an equation similar to Van Der Pol equation which describe the oscillators.

It was shown that the resonant-tunneling diode display an oscillator by using mathematica programming with Runge kutta method, also I used simulation by Orcad to show that some transmission lines reshape the signals and display oscillators.

# Appendix A

## Transmission Line Equations

Again rewrite Eqs. (1.3,1.5,1.6,1.7)

$$-\frac{\partial V(z,t)}{\partial z} = RI(z,t) + L\frac{\partial I(z,t)}{\partial t} \quad (\text{A.1})$$

$$-\frac{\partial I(z,t)}{\partial z} = GV(z,t) + C\frac{\partial V(z,t)}{\partial t} \quad (\text{A.2})$$

the solution is

$$V(z,t) = R_e[V_s(z)e^{j\omega t}] \quad (\text{A.3})$$

$$I(z,t) = R_e[I_s(z)e^{j\omega t}] \quad (\text{A.4})$$

Substituting this in Eqs.(A.1),(A.2) we get

$$-\frac{\partial(R_e[V_se^{j\omega t}])}{\partial z} = RR_e[I_se^{j\omega t}] + L\frac{\partial}{\partial t}R_e[I_se^{j\omega t}] \quad (\text{A.5})$$

$$-R_e e^{j\omega t} \frac{\partial V_s}{\partial z} = RR_e I_s e^{j\omega t} + jL\omega R_e I_s e^{j\omega t} \quad (\text{A.6})$$

dividing by  $R_e e^{j\omega t}$  we get

$$-\frac{\partial V_s}{\partial z} = RI_s + LI_s\omega j \quad (\text{A.7})$$

$$-\frac{\partial V_s}{\partial z} = (R + j\omega L)I_s \quad (\text{A.8})$$

similarly we get

$$-\frac{\partial I_s}{\partial z} = (G + j\omega C)V_s \quad (\text{A.9})$$

# Appendix B

## Characteristic Impedance

$$V_s(z) = V_o^+ e^{-\gamma z} + V_o^- e^{\gamma z} \quad (\text{B.1})$$

$$I_s(z) = I_o^+ e^{-\gamma z} + I_o^- e^{\gamma z} \quad (\text{B.2})$$

where

$V_o^+, I_o^+, V_o^-, I_o^-$ , are wave amplitude along  $+z$  and  $-z$  directions

The instantaneous voltage

$$V(z) = R_e[V_s e^{j\omega t}] \quad (\text{B.3})$$

substitute from Eqs .(B.1),(B.2)into Eq.(A.8) we get

$$\frac{-\partial}{\partial z}(V_o^+ e^{-\gamma z} + V_o^- e^{\gamma z}) = (R + \omega j l)(I_o^+ e^{-\gamma z} + I_o^- e^{\gamma z}) \quad (\text{B.4})$$

$$+\gamma V_o^+ e^{-\gamma z} - \gamma V_o^- e^{\gamma z} = (R + \omega j l)(I_o^+ e^{-\gamma z} + I_o^- e^{\gamma z}) \quad (\text{B.5})$$

equating coefficient of  $e^{\gamma z}, e^{-\gamma z}$

$$\gamma V_o^+ = (R + \omega j l) I_o^+ \quad (\text{B.6})$$

$$Z_o = \frac{V_o^+}{I_o^+} = \frac{(R + j\omega l)}{\gamma} \quad (\text{B.7})$$

similarly from current equation( A.9) we can get

$$Z_o = \frac{V_o^+}{I_o^+} = \frac{\gamma}{G + j\omega C} \quad (\text{B.8})$$



$$Z_o = \sqrt{\frac{(G + j\omega C)(R + j\omega l)}{(G + j\omega C)^2}} \quad (\text{B.9})$$

$$Z_o = \sqrt{\frac{(R + j\omega l)}{(G + j\omega C)}} \quad (\text{B.10})$$

# Appendix C

## Equations of Smith Chart

$$z_L = \frac{Z_L}{Z_o} = r + jx \quad (\text{C.1})$$

$$\Gamma = \frac{z_L - 1}{z_L + 1} \quad (\text{C.2})$$

$$z_L(1 - \Gamma) = 1 + \Gamma \quad (\text{C.3})$$

$$z_L = \frac{1 + \Gamma}{1 - \Gamma} \quad (\text{C.4})$$

$$z_L = \frac{(\Gamma_r + 1) + j\Gamma_i}{(1 - \Gamma_r) - j\Gamma_i} \quad (\text{C.5})$$

Multiply by the conjugate

$$z_L = \frac{(\Gamma_r + 1) + j\Gamma_i}{(1 - \Gamma_r) - j\Gamma_i} \times \frac{(1 - \Gamma_r) + j\Gamma_i}{(1 - \Gamma_r) + j\Gamma_i} \quad (\text{C.6})$$

$$z_L = \frac{(\Gamma_r + 1)(1 - \Gamma_r) + (\Gamma_r + 1)j\Gamma_i + j\Gamma_i(1 - \Gamma_r) - \Gamma_i^2}{(1 - \Gamma_r)^2 + (1 - \Gamma_r)j\Gamma_i - j\Gamma_i(1 - \Gamma_r) + \Gamma_i^2} \quad (\text{C.7})$$

$$z_L = \frac{\Gamma_r - \Gamma_r^2 + 1 - \Gamma_r + \Gamma_r\Gamma_i j + j\Gamma_i + j\Gamma_i - \Gamma_i\Gamma_r j - \Gamma_i^2}{(1 - \Gamma_r)^2 + j\Gamma_i - j\Gamma_r\Gamma_i - j\Gamma_i + j\Gamma_i\Gamma_r + \Gamma_i^2} \quad (\text{C.8})$$

$$z_L = \frac{1 - \Gamma_r^2 - \Gamma_i^2 + 2\Gamma_i j}{(1 - \Gamma_r)^2 + \Gamma_i^2} \quad (\text{C.9})$$

$$z_L = r + jx \quad (\text{C.10})$$

$$r = \frac{1 - \Gamma_r^2 - \Gamma_i^2}{(1 - \Gamma_r)^2 + \Gamma_i^2} \quad (\text{C.11})$$

$$x = \frac{2\Gamma_i}{(1 - \Gamma_r)^2 + \Gamma_i^2} \quad (\text{C.12})$$

From (B.11) we get

$$r(1 - \Gamma_r)^2 + r\Gamma_i^2 = 1 - \Gamma_r^2 - \Gamma_i^2 \quad (\text{C.13})$$

$$r(1 - 2\Gamma_r + \Gamma_r^2) + r\Gamma_i^2 = 1 - \Gamma_r^2 - \Gamma_i^2 \quad (\text{C.14})$$

$$r(1 - 2\Gamma_r + \Gamma_r^2) + \Gamma_r^2 - 1 = -\Gamma_i^2(1 + r) \quad (\text{C.15})$$

$$\Gamma_r^2(1 + r) - 2r\Gamma_r + r - 1 = -\Gamma_i^2(1 + r) \quad (\text{C.16})$$

$$(1 + r)[\Gamma_r^2 - \frac{2r}{1 + r}\Gamma_r] + r - 1 = -\Gamma_i^2(1 + r) \quad (\text{C.17})$$

By complete the square

$$(1 + r)[\Gamma_r^2 - \frac{2r}{1 + r}\Gamma_r + (\frac{r}{1 + r})^2] - \frac{r^2}{1 + r} + r - 1 = -\Gamma_i^2(1 + r) \quad (\text{C.18})$$

$$(\Gamma_r - \frac{r}{1 + r})^2 + \Gamma_i^2 = \frac{r^2}{(1 + r)^2} + \frac{1 - r}{1 + r} \quad (\text{C.19})$$

$$(\Gamma_r - \frac{r}{1 + r})^2 + \Gamma_i^2 = \frac{r^2 + (1 - r)(1 + r)}{(1 + r)^2} \quad (\text{C.20})$$

$$(\Gamma_r - \frac{r}{1 + r})^2 + \Gamma_i^2 = (\frac{1}{1 + r})^2 \quad (\text{C.21})$$

And this is equation of a circle of the form

$$(x - a)^2 + (y - b)^2 = r^2 \quad (\text{C.22})$$

The center

$$(\Gamma_r, \Gamma_i) = (\frac{r}{1+r}, 0) \quad (\text{C.23})$$

And the radius is

$$\frac{1}{1+r} \quad (\text{C.24})$$

From (B.12) we have

$$x = \frac{2\Gamma_i}{(1 - \Gamma_r)^2 + \Gamma_i^2} \quad (\text{C.25})$$

$$x(1 - \Gamma_r)^2 + x\Gamma_i^2 = 2\Gamma_i \quad (\text{C.26})$$

$$x(1 - 2\Gamma_r + \Gamma_r^2) + x\Gamma_i^2 = 2\Gamma_i \quad (\text{C.27})$$

$$x(\Gamma_r - 1)^2 + x[\Gamma_i^2 - \frac{2}{x}\Gamma_i] = 0 \quad (\text{C.28})$$

complete the square for the second term and dividing by x yields

$$x(\Gamma_r - 1)^2 + x[\Gamma_i^2 - \frac{2}{x}\Gamma_i + \frac{1}{x^2}] - \frac{1}{x} = 0 \quad (\text{C.29})$$

complete the square and divided by x we have

$$(\Gamma_r - 1)^2 + (\Gamma_i - \frac{1}{x})^2 = (\frac{1}{x})^2 \quad (\text{C.30})$$

And this is equation of a circle

at center  $(1, \frac{1}{x})$  and radius  $= \frac{1}{x}$

## Appendix D

# Numerical Solution For Van Der Pol Equation

```
DSolve[y''[x] - μ(1 - y[x]^2)y'[x] + y[x] == 0, y[x], x]
DSolve[y[x] - μ(1 - y[x]^2)y'[x] + y''[x] == 0, y[x], x]
Plot
soln[μ, t] := NDSolve[y''[x] - μ(1 - y[x]^2)y'[x] + y[x] == 0.
y[0] = .01, y'[0] = 0, y[x], x, 0, t][[1, 1, 2, 0]];
plot[t, μ] := Block[DisplayFunction = Identity],
ParametricPlot[Evaluate[x, soln[m, t][x]], x, 0, t,
PlotStyle -> Red, AxesLabel -> TraditionalForm/@x, y[x],
PlotLabel -> TraditionalForm[μ == m]],
ParametricPlot[Evaluate[soln[m, t][x], D[soln[m, t][x], x]], x, 0, t,
AspectRatio -> Automatic, PlotStyle -> Red,
PlotPoints -> 100, AxesLabel -> TraditionalForm/@y[x], y'[x],
PlotRange -> All]
Show[GraphicsArray[plot[100, ]/@.2, 1, 5
[, GraphicsSpacing -> -.15, 0, ImageSize -> 400]
```

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